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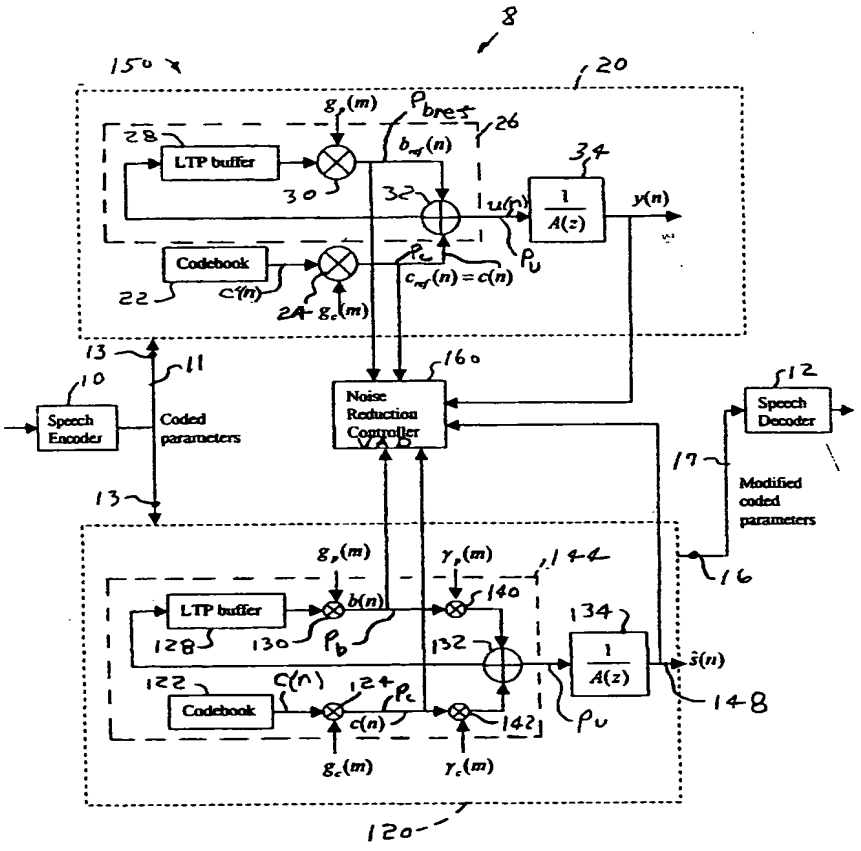

(57) Abstract: A communications system (8) transmits digital signals using a compression code comprising a plurality of parameters including a first parameter. The parameters represent an audio signal comprising a plurality of audio characteristics, including a noise characteristic. The compression code is decodable by a plurality of decoding steps. A processor (150) is responsive to the compression code to read at least the first parameter. Based on such signals, the processor adjusts the first parameter and writes the adjusted first parameter into the compression code. As a result, the noise condition is effectively managed.

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TITLE OF THE INVENTION

CODED DOMAIN NOISE CONTROL

CROSS-REFERENCE TO RELATED APPLICATIONS

This is a utility application corresponding to provisional application no. 60/142,136 entitled "CODED DOMAIN ENHANCEMENT OF COMPRESSED SPEECH " filed July 2, 1999.

BACKGROUND OF THE INVENTION

The present invention relates to coded domain enhancement of compressed speech and in particular to coded domain noise contol.

This specification will refer to the following references:

[1] GSM 06.10, "Digital cellular telecommunication system (Phase 2); Full rate speech; Part 2: Transcoding", ETS 300 580-2, March 1998.

[2] GSM 06.60, "Digital cellular telecommunications system (Phase 2); Enhanced Full Rate (EFR) speech transcoding", June 1998.

[3] GSM 08.62, "Digital cellular telecommunications system (Phase 2+); Inband Tandem Free Operation (TFO) of Speech Codecs", ETSI, March 2000.

[4] J. R. Deller, J. G. Proakis, J. H. L. Hansen, "Discrete-Time Processing of Speech Signals", Chapter 7, Prentice-Hall Inc, 1987.

[5] S. V. Vaseghi, "Advanced Signal Processing and Digital Noise Reduction", Chapter 9, Wiley (ISBN 0471958751), 1996.

The specification may refer to the following abbreviations:

ACELP	Algebraic Code Excited Linear Prediction
AE	Audio Enhancer
ALC	Adaptive or Automatic Level Control
CD	Coded Domain or Compressed Domain
CDEC	Coded Domain Echo Control
CDNR	Coded Domain Noise Reduction
EFR	Enhanced Full Rate
ETSI	European Telecommunications Standards Institute
FR	Full Rate
GSM	Global System for Mobile Communications
ITU	International Telecommunications Union
MR-ACELP	Multi-Rate ACELP

PCM	Pulse Code Modulation (ITU G.711)
RPE-LTP	Regular Pulse Excitation - Long Term Prediction
TFO	Tandem Free Operation
VSELP	Vector Sum Excitation Linear Prediction

Network enhancement of coded speech would normally require decoding, linear processing and re-encoding of the processed signal. Such a method is very expensive. Moreover, the encoding process is often an order of magnitude more computationally intensive than the speech enhancement methods.

Speech compression is increasingly used in telecommunications, especially in cellular telephony and voice over packet networks. Past network speech enhancement techniques which operate in the linear domain have several shortcomings. For example, past network speech enhancement techniques which operate in the linear domain require decoding of compressed speech, performing the necessary enhancements and re-encoding of the speech. This processing can be computationally intensive, is especially prone to additional quantization noise, and can cause additional delay.

The maintenance of the speech level at an optimal level is an important problem in the Public Switched Telephone Network (PSTN). Telephony customers expect a comfortable listening level to maximize comprehension of their conversation. The transmitted speech level from a telephone instrument depends on the speaker's volume and the position of the speaker relative to the microphone. If volume control is available on the telephone instrument, the listener could manually adjust it to a desirable level. However, for historical reasons, most telephone instruments do not have volume controls. Also, direct volume control by the listener does not address the need to maintain appropriate levels for network equipment.

Furthermore, as technology is progressing towards the era of hands-free telephony especially in the case of mobile phones in vehicles, manual adjustment is considered cumbersome and potentially hazardous to the vehicle operators.

The responsibility of maintaining speech quality has generally been the role of the network service providers, with the telephone instrument manufacturers playing a relatively minor role. Traditionally, network service providers have provided tight specifications for equipment and networks with regard to speech levels. However, due to increased international voice traffic, deregulation, fierce competition and greater customer expectations, the network service providers have to ensure the proper speech levels with lesser influence over specifications and equipment used in other networks.

With the widespread introduction of new technology and protocols such as digital cellular telephony and voice over packet networks, the control of speech levels in the network has become more complex. One of the main reasons is the presence of speech compression devices known as speech codecs (coder-decoder pairs) in the transmission path. Automatic level control (ALC) and noise reduction (NR) of speech signals becomes more difficult when speech codecs are present in the transmission path, while, in the linear domain, the digital speech samples are available for direct processing.

A need has long existed in the industry for a coded domain signal processing approach that reduces computational costs, reduces delay, and reduces additional quantization noise.

The GSM Digital Cellular Network

In the GSM digital cellular network, speech transmission between the mobile stations (handsets) and the base station is in compressed or coded form. Speech coding techniques such as the GSM FR [1] and EFR [2] are used to compress the speech. The devices used to compress speech are called *vocoders*. The coded speech requires less than 2 bits per sample. This situation is depicted in Figure 1. Between the base stations, the speech is transmitted in an uncoded form (using PCM companding which requires 8 bits per sample).

Note that the terms coded speech and uncoded speech are defined as follows:

Uncoded speech: refers to the digital speech signal samples typically used in telephony; these samples are either in linear 13-bits per sample form or companded form such as the 8-bits per sample μ-law or A-law PCM form; the typical bit-rate is 64 kbps.

Coded speech: refers to the compressed speech signal parameters (also referred to as coded parameters) which use a bit rate typically well below 64kbps such as 13 kbps in the case of the GSM FR and 12.2 kbps in the case of GSM EFR; the compression methods are more extensive than the simple PCM companding scheme; examples of compression methods are linear predictive coding, code-excited linear prediction and multi-band excitation coding.

Tandem-Free Operation (TFO) in GSM

The Tandem-Free Operation (TFO) standard [3] will be deployed in GSM digital cellular networks in the near future. The TFO standard applies to mobile-to-mobile calls. Under TFO, the speech signal is conveyed between mobiles in a compressed form after a brief negotiation period. This eliminates tandem voice codecs during mobile-to-mobile calls. The elimination of tandem codecs is known to improve speech quality in the case where the original signal is clean. The key point to note is that the speech transmission remains coded between the mobile handsets and is depicted in Figure 2.

Under TFO, the transmissions between the handsets and base stations are coded, requiring less than 2 bits per speech sample. However, 8 bits per speech sample are still available for transmission between the base stations. At the base station, the speech is decoded and then A-law companded so that 8 bits per sample are necessary. However, the original coded speech bits are used to replace the 2 least significant bits (LSBs) in each 8-bit A-law companded sample. Once TFO is established between the handsets, the base stations only send the 2 LSBs in each 8-bit sample to their respective handsets and discard the 6 MSBs. Hence vocoder tandeming is avoided. The process is illustrated in Figure 3.

The Background Noise Problem and Traditional Solutions

Environmental background noise is a major impairment that affects telephony applications. Such additive noise can be especially severe in the case of cellular telephones operated in noisy environments. Telephony service providers use noise reduction equipment in their networks to improve the quality of speech so as to encourage longer talk times and increase customer satisfaction. Although noise could be handled at the source in the case of digital cellular handsets, few handset models provide such features due to cost and power limitations. Where such features are provided, they may still not meet the service provider's requirements. Hence service providers consider network speech enhancement equipment to be essential for their competitiveness in the face of deregulation and greater customer expectations. The explosive increase in the use of cellular telephones, which are often operated in the presence of severe background noise conditions, has also increased the use of noise reduction equipment in the network.

The traditional method for noise reduction is shown in Figure 4. It is based on a well known technique called spectral subtraction [5].

In the spectral subtraction approach, the noisy signal is decomposed into different frequency bands, e.g. using the discrete Fourier transform. A silence detector is used to demarcate gaps in speech. During such silence segments, the noise spectrum (i.e. the noise power in each frequency band) is estimated. At all times, the noisy signal power in each frequency band is also estimated. These power estimates provide information such as the signal-to-noise ratio in each frequency band during

the time of measurement. Based on these power estimates, the magnitude of each frequency component is attenuated. The phase information is not changed. The resulting magnitude and phase information are recombined. Using the inverse discrete Fourier transform, a noise-reduced signal is reconstructed.

Techniques such as the one described above require the uncoded speech signal for noise reduction processing. The output of such noise reduction processing also results in an uncoded speech signal. Under TFO in GSM networks, if noise reduction is implemented in the network, a traditional approach requires decoding the coded speech, processing the resulting uncoded speech and then re-encoding it. Such decoding and re-encoding is necessary because the traditional techniques can only operate on the uncoded speech signal. This approach is shown in Figure 5. Some of the disadvantages of this approach are as follows.

This approach is computationally expensive due to the need for two decoders and an encoder. Typically, encoders are at least an order of magnitude more complex computationally than decoders. Thus, the presence of an encoder, in particular, is a major computational burden.

The delay introduced by the decoding and re-encoding processes is undesirable.

A vocoder tandem (i.e. two encoder/decoder pairs placed in series) is introduced in this approach, which is known to degrade speech quality due to quantization effects.

The proposed techniques are capable of performing noise reduction directly on the coded speech (i.e. by direct modification of the coded parameters). Low computational complexity and delay are achieved. Tandeming effects are avoided or minimized, resulting in better perceived quality after noise reduction.

Speech Coding

Overview

Speech compression, which falls under the category of lossy source coding, is commonly referred to as speech coding. Speech coding is performed to minimize the bandwidth necessary for speech transmission. This is especially important in wireless telephony where bandwidth is scarce. In the relatively bandwidth abundant packet networks, speech coding is still important to minimize network delay and jitter. This is because speech communication, unlike data, is highly intolerant of delay. Hence a smaller packet size eases the transmission through a packet network. The four ETSI GSM standards of concern are listed in Table 1.

Table 1: GSM Speech Codecs

Codec Name	Coding Method	Bit Rate (kbits/sec)
Half Rate (HR)	VSELP	5.6
Full Rate (FR)	RPE-LTP	13
Enhanced Full Rate (EFR)	ACELP	12.2
Adaptive Multi-Rate (AMR)	MR-ACELP	5.4-12.2

In speech coding, a set of consecutive digital speech samples is referred to as a speech frame. The GSM coders operate on a frame size of 20ms (160 samples at 8kHz sampling rate). Given a speech frame, a speech encoder determines a small set of parameters for a speech synthesis model. With these speech parameters and the

speech synthesis model, a speech frame can be reconstructed that appears and sounds very similar to the original speech frame. The reconstruction is performed by the speech decoder. In the GSM vocoders listed above, the encoding process is much more computationally intensive than the decoding process.

The speech parameters determined by the speech encoder depend on the speech synthesis model used. The GSM coders in Table 1 utilize linear predictive coding (LPC) models. A block diagram of a simplified view of a generic LPC speech synthesis model is shown in Figure 6. This model can be used to generate speech-like signals by specifying the model parameters appropriately. In this example speech synthesis model, the parameters include the time-varying filter coefficients, pitch periods, codebook vectors and the gain factors. The synthetic speech is generated as follows. An appropriate codebook vector, $c(n)$, is first scaled by the codebook gain factor g_c. Here n denotes sample time. The scaled codebook vector is then filtered by a pitch synthesis filter whose parameters include the pitch gain, g_p, and the pitch period, T. The result is sometimes referred to as the total excitation vector, $u(n)$. As implied by its name, the pitch synthesis filter provides the harmonic quality of voiced speech. The total excitation vector is then filtered by the LPC synthesis filter which specifies the broad spectral shape of the speech frame.

For each speech frame, the parameters are usually updated more than once. For instance, in the GSM FR and EFR coders, the codebook vector, codebook gain and the pitch synthesis filter parameters are determined every subframe (5ms). The

LPC synthesis filter parameters are determined twice per frame (every 10ms) in EFR and once per frame in FR.

Encoding Steps

Here is a summary of the typical sequence of steps used in a speech encoder:

Obtain a frame of speech samples.

Multiply the frame of samples by a window (e.g. Hamming window) and determine the autocorrelation function up to lag M .

Determine the reflection coefficients and/or LPC coefficients from the autocorrelation function. (Note that reflection coefficients are an alternative representation of the LPC coefficients.)

Transform the reflection coefficients or LPC coefficients to a different form suitable for quantization (e.g. log-area ratios or line spectral frequencies)

Quantize the transformed LPC coefficients using vector quantization techniques.

The following sequence of operations is typically performed for each subframe:

Determine the pitch period.

Determine the corresponding pitch gain.

Quantize the pitch period and pitch gain.

Inverse filter the original speech signal through the quantized LPC synthesis filter to obtain the LPC residual signal.

Inverse filter the LPC residual signal through the pitch synthesis filter to obtain the pitch residual.

Determine the best codebook vector.

Determine the best codebook gain.

Quantize the codebook gain and codebook vector.

Update the filter memories appropriately.

Add any additional error correction/detection, framing bits etc.

Transmit the coded parameters.

Decoding Steps

Here is the typical sequence of steps used in a speech decoder:

Perform any error correction/detection and framing.

For each subframe:

Dequantize all the received coded parameters (LPC coefficients, pitch period, pitch gain, codebook vector, codebook gain).

Scale the codebook vector by the codebook gain and filter it using the pitch synthesis filter to obtain the LPC excitation signal.

Filter the LPC excitation signal using the LPC synthesis filter to obtain a preliminary speech signal.

Construct a post-filter (usually based on the LPC coefficients).

Filter the preliminary speech signal to reduce quantization noise to obtain the final synthesized speech.

Arrangement of Coded Parameters in the Bit-stream

As an example of the arrangement of coded parameters in the bit-stream transmitted by the encoder, the GSM FR vocoder is considered. For the GSM FR vocoder, a frame is defined as 160 samples of speech sampled at 8kHz, i.e. a frame is 20ms long. With A-law PCM companding, 160 samples would require 1280 bits for transmission. The encoder compresses the 160 samples into 260 bits. The arrangement of the various coded parameters in the 260 bits of each frame is shown in Figure 7. The first 36 bits of each coded frame consists of the log-area ratios which correspond to LPC synthesis filter. The remaining 224 bits can be grouped into 4 subframes of 56 bits each. Within each subframe, the coded parameter bits contain the pitch synthesis filter related parameters followed by the codebook vector and gain related parameters.

Speech Synthesis Transfer Function and Typical Coded Parameters

Although many non-linearities and heuristics are involved in the speech synthesis at the decoder, the following approximate transfer function may be attributed to the synthesis process:

$$H(z) = \frac{g_c}{\left(1 - g_p z^{-T}\right)\left(1 - \sum_{k=1}^{M} a_k z^{-k}\right)} \quad (1A)$$

The codebook vector, $c(n)$, is filtered by $H(z)$ to result in the synthesized speech. The key point to note about this generic LPC model for speech decoding is that the available coded parameters that can be modified to achieve noise reduction are:

$c(n)$: codebook vector

g_c: codebook gain

g_p: pitch gain

T: pitch period

$\{a_k, k = 1,...,M\}$: LPC coefficients

Most LPC-based vocoders use parameters similar to the above set, parameters that may be converted to the above forms, or parameters that are related to the above forms. For instance, the LPC coefficients in LPC-based vocoders may be represented using log-area ratios (e.g. the GSM FR) or line spectral frequencies (e.g. GSM EFR);

both of these forms can be converted to LPC coefficients. An example of a case where a parameter is related to the above form is the block maximum parameter in the GSM FR vocoder; the block maximum can be considered to be directly proportional to the codebook gain in the model described by equation (1A).

Thus, although the discussion of coded parameter modification methods is mostly limited to the generic speech decoder model, it is relatively straightforward to tailor these methods for any LPC-based vocoder, and possibly even other models.

Applicability of Older Speech Processing Techniques to the Coded Domain

It should also be clear that techniques such as spectral subtraction used with uncoded speech for noise reduction cannot be used on the coded parameters because the coded parameter representation of the speech signal is significantly different.

BRIEF SUMMARY OF THE INVENTION

The invention is useful in a communication system for transmitting digital signals using a compression code comprising a predetermined plurality of parameters including a first parameter. The parameters represent an audio signal having a plurality of audio characteristics including a noise characteristic. The compression code is decodable by a plurality of decoding steps. In such an environment, according to one embodiment of the invention, the noise characteristic can be managed by reading at least the first parameter, and by generating an adjusted first parameter in response to the compression code and the first parameter. The first parameter is replaced with the adjusted first parameter. The reading, generating and replacing are preferably performed by a processor.

The invention also is useful in a communication system for transmitting digital signals comprising code samples further comprising first bits using a compression code and second bits using a linear code. The code samples represent an audio signal having a plurality of audio characteristics including a noise characteristic. In such an environment, according to a second embodiment of the invention, the noise characteristic can be managed without decoding the compression code by adjusting the first bits and second bits in response to the second bits.

BRIEF DESCRIPTION OF THE DRAWINGS

Figure 1 is a schematic block diagram of a system for speech transmission in a GSM digital cellular network.

Figure 2 is a schematic block diagram of a system for speech transmission in a GSM network under tandem-free operation (TFO).

Figure 3 is a graph illustrating transmission of speech under tandem-free operation (TFO).

Figure 4 is a schematic block diagram of a traditional noise reduction approach using spectral subtraction.

Figure 5 is a schematic block diagram illustrating noise reduction of coded speech using a traditional approach.

Figure 6 is a schematic block diagram of a generic LPC speech synthesis model or speech decoder model.

Figure 7 is a schematic block diagram illustrating an arrangement of coded parameters in a bit-stream for GSM FR.

Figure 8 is a schematic block diagram distinguishing coded domain digital speech parameters from linear domain digital speech samples.

Figure 9 is a graph illustrating GSM full rate codec quantization levels for block maxima.

Figure 10a is a schematic block diagram of a backward adaptive standard deviation based quantizer.

Figure 10b is a schematic block diagram of a backward adaptive differential based quantizer.

Figure 11 is a schematic block diagram of an adaptive differential quantizer using a linear predictor.

Figure 12 is a schematic block diagram of a GSM enhanced full rate codebook gain (speech level related parameter) quantizer.

Figure 13 is a graph illustrating GSM enhanced full rate codec quantization levels for a gain correction factor.

Figure 14 is a schematic block diagram of one technique for coded domain ALC.

Figure 15 is a flow diagram illustrating a technique for overflow/underflow prevention.

Figure 16 is a schematic block diagram of a preferred form of ALC system using feedback of the realized gain in ALC algorithms requiring past gain values.

Figure 17 is a schematic block diagram of one form of a coded domain ALC device.

Figure 18 is a schematic block diagram of a system for instantaneous scalar requantization for a GSM FR codec.

Figure 19 is a schematic block diagram of a system for differential scalar requantization for a GSM EFR codec.

Figure 20a is a graph showing a step in desired gain.

Figure 20b is a graph showing actual realized gain superimposed on the desired gain with a quantizer in the feedback loop.

Figure 20c is a graph showing actual realized gain superimposed on the desired gain resulting from placing a quantizer outside the feedback loop shown in Figure 19.

Figure 21 is a schematic block diagram of an ALC device showing a quantizer placed outside the feedback loop.

Figure 22 is a schematic block diagram of a simplified version of the ALC device shown in Figure 21.

Figure 23a is a schematic block diagram of a coded domain ALC implementation for ALC algorithms using feedback of past gain values with a quantizer in the feedback loop.

Figure 23b is a schematic block diagram of a coded domain ALC implementation for ALC algorithms using feedback of past gain values with a quantizer outside the feedback loop.

Figure 24 is a graph showing spacing between adjacent R_i values in an EFR codec, and more specifically showing EFR Codec SLRPs: $(R_{i+1} - R_i)$ against i.

Figure 25a is a diagram of a compressed speech frame of an EFR encoder illustrating the times at which various bits are received and the earliest possible decoding of samples as a buffer is filled from left to right.

Figure 25b is a diagram of a compressed speech frame of an FR encoder illustrating the times at which various bits are received and the earliest possible decoding of samples as a buffer is filled from left to right.

Figure 26 is a schematic block diagram illustrating a single-band linear domain noise reduction technique.

Figure 27 is a schematic block diagram of a differential scalar quantization technique.

Figure 28 is a schematic block diagram of a system for differential requentization of a differentially quantized parameter.

Figure 29 is a graph illustrating reverberations caused by differential quantization.

Figure 30 is a schematic block diagram of a system for reverberation-free differential requantization.

Figure 31 is a simplified schematic block diagram of a simplified reverberation-free differential requantization system.

Figure 32 is schematic block diagram of a dual-source view of speech synthesis.

Figure 33 is a schematic block diagram of a preferred form of network noise reduction.

Figure 34 is a graph illustrating magnitude frequency response of comb filters.

Figure 35 is a graph illustrating increase in spectral peakresponse of a comb filter due to pitch gain control.

Figure 36 is a schematic block diagram of one preferred form of a coded domain noise reduction system using codebook gain attenuation.

Figure 37 is a flow diagram of a preferred form of coded domain noise reduction methodology according to the invention.

Figure 38 is a schematic block diagram of a system for coded domain noise reduction by modification of the codebook vector parameter.

Figure 39 is a graph illustrating a spectral interpretation of line spectral frequencies.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

While the invention will be described in connection with one or more embodiments, it will be understood that the invention is not limited to those embodiments. On the contrary, the invention includes all alternatives, modifications, and equivalents as may be included within the spirit and scope of the appended claims. For example, the ALC techniques described in this specification also have application to NR techniques.

In modern networks, speech signals are digitally sampled prior to transmission. Such digital (i.e. discrete-time discrete-valued) signals are herein referred to in this specification as being in the linear domain. The adjustment of the speech levels in such linear domain signals is accomplished by multiplying every sample of the signal by an appropriate gain factor to attain the desired target speech level.

Digital speech signals that are typically carried in telephony networks usually undergo a basic form of compression such as pulse code modulation (PCM) before transmission. Such compression schemes are very inexpensive in terms of computations and delay. It is a relatively simple matter for an ALC or NR device to convert the compressed digital samples to the linear domain, process the linear samples, and then compress the processed samples before transmission. As such,

these signals can effectively be considered to be in the linear domain. In the context of this specification, compressed or coded speech will refer to speech that is compressed using advanced compression techniques that require significant computational complexity.

More specifically, in this specification and claims, linear code and compression code have the following meanings:

Linear code: By a linear code, we mean a compression technique that results in one coded parameter or coded sample for each sample of the audio signal. Examples of linear codes are PCM (A-law and μ-law) ADPCM (adaptive differential pulse code modulation), and delta modulation.

Compression code: By a compression code, we mean a technique that results in fewer than one coded parameter for each sample of the audio signal. Typically, compression codes result in a small set of coded parameters for each block or frame of audio signal samples. Examples of compression codes are linear predictive coding based vocoders such as the GSM vocoders (HR, FR, EFR).

Speech compression, which falls under the category of lossy source coding, is commonly referred to as speech coding. Speech coding is performed to minimize the bandwidth necessary for speech transmission. This is especially important in wireless telephony where bandwidth is a scarce resource. In the relatively bandwidth abundant packet networks, speech coding is still important to minimize network delay and jitter. This is because speech communication, unlike data, is highly intolerant of delay. Hence a smaller packet size eases the transmission through a packet network. Several industry standard speech codecs (coder-decoder pairs) were listed in Table 1 for reference.

In speech coding, a set of consecutive digital speech samples is referred to as a speech frame. Given a speech frame, a speech encoder determines a small set of parameters for a speech synthesis model. With these speech parameters and the speech synthesis model, a speech frame can be reconstructed that appears and sounds very similar to the original speech frame. The reconstruction is performed by the speech decoder. It should be noted that, in most speech coders, the encoding process is much more computationally intensive than the decoding process. Furthermore, the millions of instructions per second (MIPs) required to attain good quality speech coding is very high. The processing capabilities of digital signal processing chipsets have advanced sufficiently only in recent years to enable the widespread use of speech coding in applications such as cellular telephone handsets.

The speech parameters determined by the speech encoder depend on the speech synthesis model used. For instance, the coders in Table 1 utilize linear predictive coding (LPC) models. (To be more specific, these coders belong to the class of code-excited linear prediction or CELP coders.) A block diagram of a simplified view of the LPC speech synthesis model is shown in Figure 6. This model can be used to generate speech-like signals by specifying the model parameters appropriately. In this example speech synthesis model, the parameters include the time-varying filter coefficients, pitch periods, excitation vectors and gain factors. Basically, the excitation vector, c(n), is first scaled by the gain factor, G. The result is then filtered by a pitch synthesis filter whose parameters include the pitch gain, g_p, and the pitch period, T, to obtain the total excitation vector, u(n). This is then filtered by the LPC synthesis filter. Other models such as the multiband excitation model are

also used in speech coding. In this context, it suffices to note that the speech parameters together with the assumed model provide a means to remove the redundancies in the digital speech signal so as to achieve compression.

As shown in Figure 6, the overall DC gain is provided by G and ALC would primarily involve modifying G. Furthermore, the gain factor g_p may be modified to obtain a certain degree of noise reduction, if desired, in the case of noisy speech.

Among the speech parameters that are generated each frame by a typical speech encoder, some parameters are concerned with the spectral and/or waveform shapes of the speech signal for that frame. These parameters typically include the LPC coefficients and the pitch information in the case of the LPC speech synthesis model. In addition to these parameters that provide spectral information, there are usually parameters that are directly related to the power or energy of the speech frame. These speech level related parameters (SLRPs) are the key to performing ALC of coded speech. Several examples of such SLRPs will be provided below.

The first three GSM codecs in Table 1 will now be discussed. All of the first three coders process speech sampled at 8kHz and assume that the samples are obtained as 13-bit linear PCM values. The frame length is 160 samples (20ms). Furthermore, they divide each frame into four subframes of 40 samples each. The SLRPs for these codecs are listed in Table 2.

Table 2. Speech Level Related Parameters in GSM Speech Codecs

Codec Name	SLRP	Description
GSM Half Rate	$R(0)$	$R(0)$ is the average signal power of the speech frame. The signal power is computed using an analysis window which is centered over the last 100 samples of the frame. The signal power in decibels is quantized to 32 levels which are spaced uniformly in 2dB steps.
GSM Full Rate	x_{max}	x_{max} is the maximum absolute value of the elements in the subframe excitation vector. x_{max} is also termed the block maximum. All the other subframe excitation elements are normalized and then quantized with respect to this maximum. The maximum is quantized to 64 levels non-uniformly.
GSM Enhanced Full Rate	γ_{gc}	γ_{gc} is the gain correction factor between a gain factor, g_c, used to scale the subframe excitation vector and a gain factor, g_c', that is predicted using a moving average model, i.e. $\gamma_{gc} = g_c / g_c'$. The correction factor is quantized to 32 levels non-uniformly.

Depending on coder, the SLRP may be specified each subframe (e.g. the GSM FR and EFR codecs) or once per frame (e.g. the GSM HR codec).

Throughout this specification, the same variable with and without a caret above it will be used to denote the unquantized and quantized values that it holds, e.g. γ_{gc} and $\hat{\gamma}_{gc}$ and are the unquantized and quantized gain correction factors in the GSM EFR standard. Note that only the quantized SLRP, $\hat{\gamma}_{gc}$, will be available at the ALC device.

The quantized and corresponding unquantized parameters are related through the quantization function, Q(.), e.g. $\hat{\gamma}_{gc} = Q(\gamma_{gc})$. We use the notation somewhat

liberally to include not just this transformation but, depending on the context, the determination of the index of the quantized value using a look-up table or formula.

The quantization function is a many-to-one transformation and is not invertible. However, we use the 'inverse' quantization function, $Q^{-1}(.)$, to denote the conversion of a given index to it corresponding quantized value using the appropriate look-up table or formula.

Figure 8 distinguishes the coded domain from the linear domain. In the linear domain, the digital speech samples are directly available for processing. The coded domain refers to the output of speech encoders or the input of the speech decoders, which should be identical if there are no channel errors. In this context, the coded domain includes both the speech parameters and the methods used to quantize or dequantize these parameters. The speech parameters that are determined by the encoder undergo a quantization process prior to transmission. This quantization is critical to achieving bit rates lower than that required by the original digital speech signal. The quantization process often involves the use of look-up tables. Furthermore, different speech parameters may be quantized using different techniques.

Processing of speech in the coded domain involves directly modifying the quantized speech parameters to a different set of quantized values allowed by the quantizer for each of the parameters. In the case of ALC, the parameters being modified are the SLRPs. For other applications, such as noise reduction (NR), other parameters may be used.

The quantization of a single speech parameter is termed scalar quantization. When a set of parameters are quantized together, the process is called vector quantization. Vector quantization is usually applied to a set of parameters that are related to each other in some way, such as the LPC coefficients. Scalar quantization is generally applied to a parameter that is relatively independent of the other parameters. A mixture of both types of quantization methods is also possible. As the SLRPs are usually scalar quantized, focus is placed on the most commonly used scalar quantization techniques.

When a parameter is quantized instantaneously, the quantization process is independent of the past and future values of the parameter. Only the current value of the parameter is used in the quantization process. The parameter to be quantized is compared to a set of permitted quantization levels. The quantization level that best matches the given parameter in terms of some closeness measure is chosen to represent that parameter. Usually, the permitted quantization levels are stored in a look-up table at both the encoder and the decoder. The index into the table of the chosen quantization level is transmitted by the encoder to the decoder. Alternatively, given an index, the quantization level may be determined using a mathematical formula. The quantization levels are usually spaced non-uniformly in the case of SLRPs. For instance, the block maxima, x_{max}, in the GSM FR codec which has a range [0,32767] is quantized to the 64 levels shown in Figure 9. In this quantization scheme, the level that is closest but higher than x_{max} is chosen. Note that the vertical axis which represents the quantization levels is plotted on a logarithmic scale.

Instantaneous quantization schemes suffer from higher quantization errors due to the use of a fixed dynamic range. Thus, adaptive quantizers are often used in speech coding to minimize the quantization error at the cost of greater computational complexity. Adaptive quantizers may utilize forward adaptation or backward adaptation. In forward adaptation schemes, extra side information regarding the dynamic range has to be transmitted periodically to the decoder in addition to the quantization table index. Thus, such schemes are usually not used in speech coders. Backward adaptive quantizers are preferred because they do not require transmission of any side information. Two general types of backward adaptive quantizers are commonly used: standard deviation based and differential. These are depicted in Figure 10.

In the standard deviation based quantization scheme of Figure 10(a), the standard deviation of previous parameter values are used to determine a normalization factor for the current parameter value, ((n). The normalization factor divides prior to quantization. This normalization procedure allows the quantization function, Q(.), to be designed for unit variance. The look-up table index of the normalized and quantized value, $\hat{\gamma}_{norm}$ (n), is transmitted to the dequantizer where the inverse process is performed. In order for the normalization and denormalization processes to be compatible, a quantized version of the normalization factor is used at both the quantizer and dequantizer. In some variations of this scheme, decisions to expand or compress the quantization intervals may be based simply on the previous parameter input only.

In the backward adaptive differential quantization scheme of Figure 10(b), the correlation between current and previous parameter values is used to advantage. When the correlation is high, a significant reduction in the quantization dynamic range can be achieved by quantizing the prediction error, r(n). The prediction error is the difference between the actual and predicted parameter values. The same predictor for ((n) must be used at both the quantizer and the dequantizer. A linear predictor, P(z), which has the following form is usually used:

$$P(z) = \sum_{k=1}^{p} b'_k \, z^{-k} \qquad (1)$$

It can be shown readily that the differential quantization scheme can also be represented as in Figure 10 when a linear predictor, P(z), is used. Note that if we approximate the transfer function P(z)/[1-P(z)] by the linear predictor, $P_1(z)=\sum_{k=1}^{p} b_k z^{-k}$, then a simpler implementation can be achieved. This simpler differential technique is used in the GSM EFR codec for the quantization of a function of the gain correction factor, γ_{gc}. In this codec, a fourth order linear predictor with fixed coefficients, [b1,b2,b3,b4] = [0.68, 0.58, 0.34, 0.19], is used at both the encoder and the decoder.

In the EFR codec, $g_c(n)$ denotes the gain factor that is used to scale the excitation vector at subframe n. This gain factor determines the overall signal level. The quantization of this parameter utilizes the scheme shown in Figure 11 but is rather indirect. The actual 'gain' parameter that is transmitted is actually a correction

factor between $g_c(n)$ and the predicted gain, $g_c'(n)$. The correction factor, $\gamma_{gc}(n)$, defined as

$$\gamma_{gc}(n) = \frac{g_c(n)}{g_c'(n)} \tag{2}$$

is considered the actual SLRP because it is the only parameter related to the overall speech level that is accessible directly in the coded domain.

At the encoder, once the best $g_c(n)$ for the current subframe n is determined, it is divided by the predicted gain to obtain $\gamma_{gc}(n)$. The predicted gain is given by

$$g_c'(n) = 10^{0.05[\tilde{E}(n)-E_1(n)+\overline{E}]} \tag{3}$$

A 32-level non-uniform quantization is performed on $\gamma_{gc}(n)$ to obtain $\hat{\gamma}_{gc}(n)$. The corresponding look-up table index is transmitted to the decoder. In equation (3), $\overline{E}$ is a constant, $E_1(n)$ depends only on the subframe excitation vector, and $\tilde{E}(n)$ depends only on the previously quantized correction factors. The decoder, thus, can obtain the predicted gain in the same manner as the encoder using (3) once the current subframe excitation vector is received. On receipt of the correction factor $\hat{\gamma}_{gc}(n)$, the quantized gain factor can be computed as $\hat{g}_c(n) = \hat{\gamma}_{gc}(n) g_c'(n)$ using the definition in equation (2).

The quantization of the SLRP, γ_{gc}, is illustrated in Figure 12. In this Figure, $R(n)$ denotes the prediction error given by

$$R(n) = E(n) - \tilde{E}(n) = 20 \log \gamma_{gc}(n) \tag{4}$$

Note that the actual information transmitted from the encoder to the decoder are the bits representing the look-up table index of the quantized $R(n)$ parameter, $\hat{R}(n)$. This detail is omitted in Figure 12 for simplicity. Since the preferred ALC technique does not affect the channel bit error rate, it is assumed that the transmitted and received parameters are identical. This assumption is valid because the result of undetected or uncorrected errors will result in noisier decoded speech regardless of whether ALC is performed.

The quantization of the SLRP at the encoder is performed indirectly by using the mean-removed excitation vector energy each subframe. $E(n)$ denotes the mean-removed excitation vector energy (in dB) at subframe n and is given by

$$\begin{aligned} E(n) &= 10 \log\left(\frac{1}{N} g_c^2 \sum_{i=0}^{N-1} C^2(i)\right) - \overline{E} \\ &= 20 \log gc + 10 \log\left(\frac{1}{N} \sum_{i=0}^{N-1} C^2(i)\right) - \overline{E} \end{aligned} \tag{5}$$

Here N = 40 is the subframe length and $\overline{E}$ is constant. The middle term in the second line of equation (5) is the mean excitation vector energy, $E_i(n)$, i.e.

$$E_i(n) = 10 \log\left(\frac{1}{N} \sum_{i=0}^{N-1} C^2(i)\right) \tag{6}$$

The excitation vector $\{c(i)\}$ is preferred at the decoder prior to the determination of the SLRP. Note that the decoding of the excitation vector is independent of the decoding of the SLRP. It is seen that E(n) is a function of the gain factor, g_c. The quantization of $\gamma_{gc}(n)$ to $\hat{\gamma}_{gc}(n)$ indirectly causes the quantization of g_c to $\hat{g}_c$. This quantized gain factor is used to scale the excitation vector, hence setting the overall level of the signal synthesized at the decoder. is the predicted energy given by

$$\tilde{E}(n) = \sum_{i=1}^{4} b_i \hat{R}(n-i) \qquad (7)$$

where $\{\hat{R}(n-i)\}$ are previously quantized values.

The preferred method of decoding the gain factor,__, will now be discussed. First, the decoder decodes the excitation vector and computes $E_I(n)$ using equation (6). Second, the predicted energy is computed using previously decoded gain correction factors using equation (7). Then the predicted gain, $g^1(c)$, is computed using equation (3). Next, the received index of the correction factor for the current subframe is used to obtain from the look-up table. Finally, the quantized gain factor is obtained as $\hat{g}_c(n) = \hat{\gamma}_{gc}(n) g_c^1(n)$. The 32 quantization levels for are illustrated in Figure 13. Note that the vertical axis in Figure 13 which represents the quantization levels is plotted on a logarithmic scale.

Those skilled in communications recognize that the quantizer techniques descriged in connection with SLRPs apply equally to NR parameters.

For most codecs, only a partial decoding of the coded speech is necessary to perform ALC. The speech is decoded to the extent necessary to extract the SLRP as well as other parameters essential for obtaining sufficiently accurate speech level, voice activity and double-talk measurements. Some examples of situations where only partial decoding suffices include:

1) In CELP decoders, a post-filtering process is performed on the signal decoded using the LPC-based model. This post-filtering helps to reduce quantization noise but does not change the overall power level of the signal. Thus, in partial decoding of CELP-coded speech, the post-filtering process can be avoided for economy.

2) Some form of silence suppression scheme is often used in cellular telephony and voice over packet networks. In these schemes, coded speech frames are transmitted only during voice activity and very little transmission is performed during silence. The decoders automatically insert some comfort noise during the silence periods to mimic the background noise from the other end. One example of such a scheme used in GSM cellular networks is called discontinuous transmission (DTX). By monitoring the side information that indicates silence suppression, the decoder in the ALC device can completely avoid decoding the signal during silence. In such cases, the determination of voice and double-talk activities can also be simplified in the ALC device.

3) In the proposed Tandem-Free Operation (TFO) standard for speech codecs in GSM networks, the coded speech bits for each channel will be carried through the

wireline network between base stations at 64 kbits/sec. This bitstream can be divided into 8-bit samples. The 2 least significant bits of each sample will contain the coded speech bits while the upper 6 bits will contain the bits corresponding to the appropriate PCM samples. The conversion of the PCM information to linear speech is very inexpensive and provides a somewhat noisy version of the linear speech signal. It is possible to use this noisy linear domain speech signal to perform the necessary voice activity, double-talk and speech level measurements as is usually done in linear domain ALC algorithms. Thus, in this case, only a minimal amount of interpretation of the PCM samples is necessary. The SLRP and any other parameters that are required for the requantization of the SLRP would have to be interpreted. The other parameters would be decoded only to the extent necessary for requantization of the SLRP. This will be clear from the examples that will follow in later sections.

Thus, we see that it is possible to implement an ALC device that only performs partial decoding and re-encoding, hence minimizing complexity and reducing quantization noise. However, the ALC approach illustrated in Figure 14 can be improved. The sub-optimality is due to the implicit assumption that the process of gain determination is independent of SLRP requantization. In general, this assumption may not be valid.

Those skilled in communications recognize that the ALC approach shown in Figure 14 also is applicable to NR.

There are three main factors which suggest an improvement over the Figure 14 approach. First, note that requantization results in a realized SLRP that usually differs

from the desired value. Hence the desired gain that was applied by the Gain Determination block will differ from the gain that will be realized when the signal is decoded. When decoding, overflow or underflow problems may arise due to this difference because the speech signal may be over-amplified or over-suppressed, respectively. Second, some ALC algorithms may utilize the past desired gain values to determine current and future desired gain values. Since the desired gain values do not reflect the actual realized gain values, such algorithms may perform erroneously when applied as shown in Figure 14. Third, the requantization process can sometimes result in undesirable reverberations in the SLRP. This can cause the speech level to be modulated unintentionally, resulting in a distorted speech signal. Such SLRP reverberations are encountered in feedback quantization schemes such as differential quantization.

Turning now to Figure 15, to overcome the overflow/underflow problems, the iterative techniques of Figure 15 can be incorporated in the Gain Determination block. Basically, after deciding on a desired gain value, the realized gain value after requantization of the SLRP may be computed. The realized gain is checked to see if overflow or underflow problems could occur. This could be accomplished, for example, by determining what the new speech level would be by multiplying the realized gain by the original speech level. Alternatively, a speech decoder could be used in the ALC device to see whether overflow/underflow actually occurs. Either way, if the realized gain value is deemed to be too high or too low, the new SLRP is reduced or increased, respectively, until the danger of overflow/underflow is considered to be no longer present.

In ALC algorithms where past desired gain values are fed back into the algorithm to determine current and future gain values, the following modification may be made. Basically, the gain that is fed back should be the realized gain after the SLRP requantization process, not the desired gain. A preferred approach is shown in Figure 16. If the desired gain was used in the feedback loop instead of the realized gain, the controller would not be tracking the actual decoded speech signal level, resulting in erroneous level control.

Note that the iterative scheme for overflow/underflow prevention of Figure 15 may also be incorporated into the Gain Determination block of Figure 16.

Finally, the methods to avoid SLRP reverberations in feedback-based quantization schemes will be discussed in detail below. In general, these methods preferably include the integration of the gain determination and SLRP requantization techniques.

Hence the joint design and implementation of the Gain Determination block and SLRP Requantization block is preferred to prevent overflow and underflow problems during decoding, ensure proper tracking by feedback-based ALC systems, and avoid the oscillatory effects introduced by feedback quantization schemes. Figure 17 illustrates the general configuration of an ALC device that uses joint gain determination and SLRP requantization. The details will depend on the particular ALC device.

The techniques for requantization of SLRPs will now be discussed. In most speech encoders, the quantization of the SLRP is performed using either instantaneous

scalar quantization or differential scalar quantization, which were discussed above. The requantization of the SLRPs for these particular cases will be described while noting that the approaches may be easily extended to any other quantization scheme. The joint determination of the gain and SLRP requantization in the ALC device configuration of Figure 17 may utilize the requantization techniques described here.

The original value of the quantized SLRP will be denoted by $\hat{\gamma}(n)$, where n is the frame or subframe index. The set of m quantization table values will be denoted by $\{\hat{\gamma}_1 \ldots \hat{\gamma}_m\}$. Depending on the speech coder, these values may, instead, be defined using a mathematical formula. The desired gain determined by the ALC device will be denoted by $g(n)$. The realized gain after SLRP requantization will be denoted by $\hat{g}(n)$. In instantaneous scalar requantization, the goal is to minimize the difference between $g(n)$ and $\hat{g}(n)$. The basic approach involves the selection of the quantization table index, k, as

$$k = \arg\min_i \| lg(n)\hat{\gamma}(n) - \hat{\gamma}_i \| \quad (8)$$

The requantized SLRP is then given by $\hat{\gamma}_{alc}(n) = \hat{\gamma}_k$.

If overflow and underflow prevention are desired, then the iterative scheme described in Figure 15 may be used. In another approach for overflow/underflow prevention, the partial decoding of the speech samples using the requantized SLRP may be performed to the extent necessary. This, of course, involves additional complexity in the algorithm. The decoded samples can then be directly inspected to ensure that overflow or underflow has not taken place.

Note that for a given received $\hat{y}(n)$, there are m possible realized gain values. For each quantization table value, all the realized gains can be precomputed and stored. This would require the storage of m^2 realized gain values, which is often feasible since m is usually a small power of two, e.g. m = 32 in the GSM EFR codec and $m = 64$ in the GSM FR codec.

If the SLRP quantization table values are uniformly spaced (either linearly or logarithmically), then it is possible to simplify the scalar requantization process. This simplification is achieved by allowing only a discrete set of desired gain values in the ALC device. These desired gain values preferably have the same spacing as the SLRP quantization values, with 0dB being one of the gains. This ensures that the desired and realized gain values will always be aligned so that equation (8) would not have to be evaluated for each table value. Hence the requantization is greatly simplified. The original quantization index of the SLRP is simply increased or decreased by a value corresponding to the desired gain value divided by the SLRP quantization table spacing. For instance, suppose that the SLRP quantization table spacing is denoted by Δ. Then the discrete set of permitted desired gain values would be 1+{..., -2, -, 0, , 2, ...} if the SLRP quantization table values are uniformly spaced linearly, and 0+{..., -2, -, 0, , 2, ...} if the SLRP quantization table values are uniformly spaced logarithmically. If the desired gain value was $1+ k_1\Delta$ (linear case) or $k_1\Delta$ (logarithmic case), then the index of the requantized SLRP is simply obtained by adding k_1 to the original quantization index of the SLRP.

Note that this low complexity instantaneous scalar requantization technique can be applied even if the SLRP quantization table values are not uniformly spaced.

In this case, Δ would be the average spacing between adjacent quantization table values, where the average is performed appropriately using either linear or logarithmic distances between the values.

An example of instantaneous scalar requantization is shown for the GSM FR codec in Figure 18. This codec's SLRP is the block maximum, x_{max}, which is transmitted every subframe. The Q and Q^{-1} blocks represent the SLRP requantization and dequantization, respectively. The index of the block maximum is first dequantized using the look-up table to obtain x_{max}. Then, x^{max} is multiplied by the desired gain to obtain $x_{max,ALC}$ which is then requantized by using the look-up table. The index of the requantized x_{max} is then substituted for the original value in the bitstream before being sent out. This requantization technique forms the basic component of all the techniques described in Figures 14-17 when implementing coded domain ALC for the GSM FR standard.

Application of the above technique to SLRPs will now be discussed, although the techniques will be applicable to other parameters just as well, such as NR related parameters. The GSM EFR codec will be used as an example for illustrating the implementation of coded domain ALC using this requantization technique.

Figure 19 shows a general coded domain ALC technique with only the components relevant to ALC being shown. Note that (G(n) denotes the original *logarithmic* gain value determined by the encoder. In the case of the EFR codec, G(n) is equal to E(n) defined in equation (5) and R(n) is as defined in equation (4). The ALC device

determines the desired gain, $\Delta^G(n)$. The SLRP, $\hat{R}(n)$, is modified by the ALC device to $\hat{R}_{ALC}(n)$ based on the desired gain. The realized gain, $\Delta\hat{R}(n)$, is the difference between original and modified SLRPs, i.e.

$$\Delta\hat{R}(n) = \hat{R}_{ale}(n) - \hat{R}(n) \tag{9}$$

Note that this is different from the actual gain realized at the decoder which, under steady-state conditions, is $[1 + P_1(1)]\Delta\hat{R}(n)$. To make the distinction clear, we will refer to the former as the *SLRP realized gain* and the latter as the *actual realized gain*. The actual realized gain is essentially an amplified version of the SLRP realized gain due to the decoding process, under steady-state conditions. By steady-state, it is meant that $\Delta G(n)$ is kept constant for a period of time that is sufficiently long so that $\Delta\hat{R}(n)$ is either steady or oscillates in a regular manner about a particular level.

This method for differential scalar requantization basically attempts to mimic the operation of the encoder at the ALC device. If the presence of the quantizers at the encoder and the ALC device is ignored, then both the encoder and the ALC device would be linear systems with the same transfer function, $Y[1 + P_1(3)]$, with the result that $\hat{G}_{ALC}(n) = G(n) + \Delta G(n)$. However, due to the quantizers which make these systems non-linear, this relationship is only approximate. Hence, the decoded gain given by

$$\hat{G}_{alc}(n) = G(n) + \Delta G(n) + \text{quantization error} \tag{10}$$

where $(\Delta G(n) + \text{quantization error})$ is the actual realized gain.

The feedback of the SLRP realized gain, $\Delta\hat{R}(n)$, in the ALC device can cause undesirable oscillatory effects. As an example, we will demonstrate these oscillatory effects when the GSM EFR codec is used. Recall that, for this codec, $P_1(z)$ has four delays elements. Each element could contain one of 32 possible values. Hence the non-linear system in the ALC device can be in any one of over a million possible states at any given time. This is mentioned because the behavior of this non-linear system is heavily influenced by its initial conditions.

The reverberations in the actual realized gain in response to a step in the desired gain, $\Delta G(n)$, will now be illustrated. For simplicity, it is assumed that the original SLRP, $\hat{R}(n)$, is constant over 100 subframes, and that the memory of $P_1(z)$ is initially zero. Figure 20(a) shows the step in the desired gain. Figure 20(b) shows the actual realized gain superimposed on the desired gain. Although the initial conditions and the original SLRP will determine the exact behavior, the reverberations in the actual realized gain shown here are quite typical.

The reverberations in the SLRP realized gain shown in Figure 20(b) cause a modulation of the speech signal and can result in audible distortions. Thus, depending on the ALC specifications, such reverberations may be undesirable. The reverberations can be eliminated by 'moving' the quantizer outside the feedback loop as shown in Figure 20. (In this embodiment, the computation of is unnecessary but is included for comparison to Figure 19.)

Placing the quantizer outside the feedback loop results in the actual realized gain shown in Figure 20(c), superimposed on the desired gain. It should be noted that, although reverberations are eliminated, the average error (i.e. the average difference between the desired and actual realized gains) is higher than that shown in Figure 20(b). Specifically, in these examples, the average error during steady state operation of the requantizer with and without the quantizer in the feedback loop are 0.39dB and 1.03dB, respectively.

The ALC apparatus of Figure 21 can be simplified as shown in Figure 22, resulting in savings in computation. This is done by replacing the linear system $Y[1+P_1(z)]$ with the constant, $\frac{1}{[1+P_1(1)]}$.

For the purposes of ALC, this simpler implementation is often found to be satisfactory especially when the desired gains are changed relatively infrequently. By infrequent changes, it is meant that the average number of subframes between changes is much greater than the order of $P_1(z)$.

Some ALC algorithms may utilize past gain values to determine current and future gain values. In such feedback-based ALC algorithms, the gain that is fed back should be the actual realized gain after the SLRP requantization process, not the desired gain. This was discussed above in conjunction with Figure 16.

Differential scalar requantization for such feedback-based ALC algorithms can be implemented as shown in Figure 23. In these implementations, the ALC device is mimicking the actions of the decoder to determine the actual realized gain.

If a simplified ALC device implementation similar to Figure 21 is desired in Figure 23(b), then the linear system $\frac{1}{[1+P_1(z)]}$ may be replaced with the constant multiplier, $\frac{1}{[1+P_1(1)]}$. A further simplification can be achieved in Figure 23(b) by replacing the linear system $1+P_1(z)$ with the constant multiplier $1+P_1(1)$, although accuracy in the calculation of the actual realized gain is somewhat reduced. In a similar manner, the implementation shown in Figure 23(a) can be simplified by replacing the linear system by with the constant multiplier $P_1(1)$.

In applications that are tolerant to reverberations but require higher accuracy in matching the desired and actual realized gains, any of the methods described earlier that have quantizers within the feedback loop may be used. For applications that cannot allow reverberations in the actual realized gains but can tolerate lower accuracy in matching the desired and actual realized gains, any of the methods described earlier that have quantizers outside the feedback loop may be used. If, however, both accuracy and avoidance of reverberations are necessary as is often the case in ALC, then a different approach is necessary.

The current method avoids reverberations in the actual realized gains by placing the quantizers outside the feedback loop as in Figures 21, 22, or 23(b). Additionally, the average error between desired and actual realized gains is minimized by restricting the desired gain values to belong to the set of possible actual realized gain values, given the current original SLRP value, $\hat{R}(n)$.

Let the set of m possible SLRP values be $\{R_0, R_1, R_2, ..., R_{m-1}\}$. Given the original SLRP, $\hat{R}(n)$, that is received from the encoder, the ALC device computes the set of m values, $\{ [R_i - \hat{R}(n)][1 + P_1(1)] \}$. This is the set of possible actual realized gain values. The ALC algorithm should preferably be designed such that the desired gain, , is selected from this set. Such restrictions can be easily imposed on a large variety of ALC algorithms since most of them already operate using a finite set of possible desired gain values.

If the R_i values are uniformly spaced, i.e. $R_{i+1} - R = \Delta$, the above restriction on the desired gain values is further simplified to selecting a desired gain value that is a multiple of the constant $\Delta[1 + P_1(1)]$. This reduces computations significantly as the desired gain value is independent of the current original SLRP value, $\hat{R}(n)$.

Even when the values are not uniformly spaced, such simplifications are usually possible. For instance, the 32 R_i values in the EFR codec can be divided into three sets, each with approximately uniform spacing. The spacing between adjacent R_i values is illustrated in Figure 24. Most of the values lie in the middle region and have an average spacing of 1.214dB. For this codec, $[HP_1(1)] = 2.79$. Thus the desired gain values are selected to be multiples of 1.214x2.79 = 3.387 dB when $\hat{R}(n)$ falls in the middle region. A further simplification is possible by always setting the desired gain value to be a multiple of 3.387dB regardless of for this codec. This is because $\hat{R}(n)$ will fall into the lower or higher regions only for very short durations

such as at the transitions between speech and silence. Hence reverberations cannot be sustained in these regions.

Thus, in general, for each uniformly spaced subset of possible SLRP values with a spacing Δ, the desired gain value can be selected to be a multiple of $\Delta[1 + P_1(1)]$ if the corresponding current original SLRP belongs to that subset.

Large buffering, processing and transmission delays are already incurred by speech coders. Further processing of the coded speech for speech enhancement purposes can add additional delay. Such additional delay is undesirable as it can potentially make telephone conversations less natural. Furthermore, additional delay may reduce the effectiveness of echo cancellation at the handsets, or alternatively, increase the necessary complexity of the echo cancellers for a given level of performance. It should be noted that implementation of ALC in the linear domain will always add at least a frame of delay due to the buffering and processing requirements for decoding and re-encoding. For the codecs listed in Table 1, note that each frame is 20ms long. However, coded domain ALC can be performed with a buffering delay much less than one frame. Those skilled in communications recognize that the same principles apply to NR.

The EFR encoder compresses a 20ms speech frame into 244 bits. At the decoder in the ALC device, the earliest point at which the first sample can be decoded is after the reception of bit 91 as shown in Figure 25(a). This represents a buffering delay of approximately 7.46ms. It turns out that sufficient information is received to decode not just the first sample but the entire first subframe at this point. Similarly,

the entire first subframe can be decoded after about 7.11ms of buffering delay in the FR decoder.

The remaining subframes, for both coders, require shorter waiting times prior to decoding. Note that each subframe has an associated SLRP in both the EFR and FR coding schemes. This is generally true for most other codecs where the encoder operates at a subframe level.

From the above, it can be realized that ALC and NR in the coded domain can be performed subframe-by-subframe rather than frame-by-frame. As soon as a subframe is decoded and the necessary level measurements are updated, the new SLRP computed by the ALC device can replace the original SLRP in the received bitstream.

The delay incurred before the SLRP can be decoded is determined by the position of the bits corresponding to the SLRP in the received bitstream. In the case of the FR and EFR codecs, the position of the SLRP bits for the first subframe determines this delay.

Most ALC algorithms determine the gain for a speech sample only after receiving that sample. This allows the ALC algorithm to ensure that the speech signal does not get clipped due to too large a gain, or underflow due to very low gains. However, in a robust ALC algorithm, both overflow and underflow are events that have low likelihoods. As such, one can actually determine gains for samples based on information derived only from previous samples. This concept is used to achieve

near-zero buffering delay in coded domain ALC for some speech codecs. Those skilled in communications recognize that the same principles apply to NR algorithms.

Basically, the ALC algorithm must be designed to determine the gain for the current subframe based on previous subframes only. In this way, almost no buffering delay will be necessary to modify the SLRP. As soon as the bits corresponding to the SLRP in a given subframe are received, they will first be decoded. Then the new SLRP will be computed based on the original SLRP and information from the previous subframes only. The original SLRP bits will be replaced with the new SLRP bits. There is no need to wait until all the bits necessary to decode the current subframe are received. Hence, the buffering delay incurred by the algorithm will depend on the processing delay which is small. Information about the speech level is derived from the current subframe only after replacement of the SLRP for the current subframe. Those skilled in communications recognize that the same principles apply to NR algorithms.

Note that most ALC algorithms can be easily converted to operate in this delayed fashion. Although there is a small risk of overflow or underflow, such risk will be isolated to only a subframe (usually about 5ms) of speech. For instance, after overflow in a subframe due to a large gain being applied, the SLRP computed for the next subframe can be appropriately set to minimize the likelihood of continued overflows. Those skilled in communications recognize that the same principles apply to NR algorithms.

This near-zero buffering delay method is especially applicable to the FR codec since the decoding of the SLRP for this codec does not involve decoding any other parameters. In the case of the EFR codec, the subframe excitation vector is also needed to decode the SLRP and the more complex differential requantization techniques have to be used for requantizing the SLRP. Even in this case, significant reduction in the delay is attained by performing the speech level update based on the current subframe after the SLRP is replaced for the current subframe. Those skilled in communications recognize that the same principles apply to NR.

Performing coded domain ALC in conjunction with the proposed TFO standard in GSM networks was discussed above. According to this standard, the received bitstream can be divided into 8-bit samples. The 2 least significant bits of each sample will contain the coded speech bits while the upper 6 bits will contain the bits corresponding to the appropriate PCM samples. Hence a noisy version of the linear speech samples is available to the ALC device in this case. It is possible to use this noisy linear domain speech signal to perform the necessary voice activity, double-talk and speech level measurements as is usually done in linear domain ALC algorithms. Thus, in this case, only a minimal amount of decoding of the coded domain speech parameters is necessary. Only parameters that are required for the determination and requantization of the SLRP would have to be decoded. Partial decoding of the speech signal is unnecessary as the noisy linear domain speech samples can be relied upon to measure the speech level as well as perform voice activity and double-talk detection.

An object of the present invention is to derive methods to perform noise reduction in the coded domain via methods that are less computationally intensive than using linear domain techniques of similar quality that require re-encoding of the processed signal. The flexibility available in the coded domain to modify parameters to effect desired changes in the signal characteristics may be limited due to quantization. A survey of the different speech parameters and the corresponding quantization methods used by industry standard speech coders was performed. The modification of the different speech parameters will be considered, in turn, and possible methods for utilizing them to achieve noise reduction will be discussed.

Due to the non-stationary nature of speech, 'short-time' measurements are preferably used to obtain information about the speech at any given time. For instance, the short-time power or energy of a speech signal is a useful means for inferring the amplitude variations of the signal. A preferred method utilizes a recursive averaging technique. In this technique, the short-time power, P(n), of a discrete-time signal s(n) is defined as

$$P(n) = BP(n-1) + \alpha 5^2(n)$$

(11)

The transfer function, $H_1(z)$, of this recursive averaging filter that has $5^2(n)$ as its input and P(n) as its output is

$$H_p(z) = \frac{\alpha}{1 - B_{z^{-1}}}, B\angle I$$

(12)

Note that the DC gain of this filter is $H_p(1) = \frac{\alpha}{(1-B)}$. This IIR filter has a pole at which can be thought of as a forgetting factor. The closer β is to unity, the slower the short-time power changes. Thus, the rate at which the power of newer samples is incorporated into the power measure can be controlled through β. The DC gain parameter α is usually set to $1-\beta$ for convenience to obtain a unity gain filter.

In some circumstances, the root-mean-square (RMS) short-time power may be more desirable. For cost-effective implementations in digital signal processors, the square-root operation is avoided by using an approximation to the RMS power by averaging the magnitude of s(n) rather than its square as follows:

$$P(n) = \beta P(n-1) + \alpha|s(n)| \quad (13)$$

If the resulting infinite length window of recursive averaging is not desirable, the power in an analysis window of size N may, for example, be averaged as follows:

$$P(n) = \frac{1}{N}\sum_{k=0}^{N-1} S^2(n) \quad (14)$$

VAD algorithms are essential for many speech processing applications. A wide variety of VAD methods have been developed. Distinguishing speech from background noise relies on the a few basic assumptions about speech. Most VAD algorithms make use of some or all of these assumptions in different ways to distinguish between speech and silence or background noise.

The first assumption is that the speech signal level is usually greater than the background noise level. This is often the most important criterion used and many

VAD algorithms are based solely on this assumption. Using this assumption, the presence of speech can be detected by comparing signal power measurements to thresholds values.

A second assumption is that speech is non-stationary while noise is relatively stationary. Using this assumption, many schemes can be devised based on steadiness of the signal spectrum or the amount of variation in the signal pitch measurements.

The development of VAD algorithms is outside the scope of this specification. Many sophisticated and robust algorithms are already available and can be applied directly on the decoded speech. As such, we will assume that, where necessary, that a good knowledge of the demarcations between speech and background noise is available.

A single-band noise reduction system is the most basic noise reduction system conceivable. In the method illustrated in Figure 26, two short-time power measurements, $P_T(u)$ and $P_N(n)$, are performed. The former is called the total power and is the sum of the speech and background noise power. The latter is the noise power. Both power measures may be performed using recursive averaging filters as given in equation (11). The total power measure is continuously updated. The noise power measure is updated only during the absence of speech as determined by the VAD. Note that the clean speech power, $P_s(n)$, can be estimated at any time as

$$P_s(n) = P_T(n) - P_N(n) \tag{15}$$

Ideally, the noise suppression is effected by a gain, $g^{(n)}$, given by

$$g(n) = \sqrt{\frac{P_s(n)}{P_T(n)}}$$

(16)

By using equation (16), the proportion of the noisy signal, y(n), that is retained after attenuation has the approximately the same power as the clean speech signal. If the signal contained temporarily contained only noise, the gain would be reduced to zero. At the other extreme, if no noise is present, then the gain would be unity. In this example, an estimate, s(n), of the clean speech signal is obtained.

In practice, note that equation (15) may actually result in a negative value for the desired signal power due to estimation errors. To avoid such a result, additional heuristics are used to ensure that is always non-negative.

A serious blemish associated with the single-band noise suppression technique is the problem of noise modulation by the speech signal. When speech is absent, the noise may be totally suppressed. However, noise can be heard at every speech burst. Hence the effect is that the noise follows the speech and the amount of noise is roughly proportional to the loudness of the speech burst. This annoying artifact can be overcome to a limited extent (but not eliminated) by limiting the lowest possible gain to a small but non-zero value such as 0.1. The modulation of the noise may be less annoying with this solution.

Among all the parameters considered, the pitch gain, g_r, and codebook gain, g_c, are perhaps the most amenable to straightforward modification. These gain parameters are relatively independent of the other parameters and are usually quantized separately. Furthermore, they usually have a good range of quantized

values (unlike the codebook excitation). The preferred embodiment uses these two parameters to achieve noise reduction.

As discussed above, the computational cost of re-encoding necessary for coded domain noise reduction can be several orders of magnitude lower than full encoding. This is true if only the pitch and codebook gains have to be requantized. The requantization process often involves searching through a table of quantized gain values and finding the value that minimizes the squared distance. A slightly more complex situation arises when a gain parameter (or any other parameter to be modified) is quantized using a differential scalar quantization scheme. Even in this case, the cost of such re-encoding is still usually several orders of magnitude lower. Requantization for a differentially quantized parameter will now be discussed.

The quantization of a single speech parameter is termed scalar quantization. When a set of parameters are quantized together, the process is called vector quantization. Vector quantization is usually applied to a set of parameters that are related to each other in some way such as the LPC coefficients. Scalar quantization is generally applied to a parameter that is relatively independent of the other parameters such as g_r, g_c and T. A mixture of both types of quantization is also possible.

When a parameter is quantized instantaneously, the quantization process is independent of the past and future values of the parameter. Only the current value of the parameter is used in the quantization process. The parameter to be quantized is compared to a set of permitted quantization levels. The quantization level that best matches the given parameter in terms of some closeness measure is chosen to represent that parameter. Usually, the permitted quantization levels are stored in a

look-up table at both the encoder and the decoder. The index into the table of the chosen quantization level is transmitted by the encoder to the decoder.

The use of instantaneous quantization schemes suffers from higher quantization errors due to the fixed dynamic range. Thus, adaptive quantizers are often used in speech coding to minimize the quantization error at the cost of greater computational complexity. A commonly used adaptive scalar quantization technique is differential quantization and a typical implementation in speech coders is illustrated in Figure 27. In a system implemented according to Figure 27, the correlation between current and previous parameter values is used to advantage. When the correlation is high, a significant reduction in the quantization dynamic range can be achieved by quantizing the prediction error, r(n). The quantized prediction error is denoted by $\hat{r}(n)$. The prediction error is the difference between the actual (unquantized) parameter, ((n), and the predicted parameter, $\gamma_{pred}(n)$. The prediction is performed using a linear predictor $P(z) = \sum_{k=1}^{P} b_{k} z^{-k}$. The same predictor for ((n) is preferably used at both the quantizer and the dequantizer. Usually, when coding speech parameters using this technique, the predictor coefficients are kept constant to obviate the need to transmit any changes to the decoder. Parameters that change sufficiently slowly such as the pitch period and gain parameters are amenable to differential quantization.

Vector quantization involves the joint quantization of a set of parameters. In its simplest form, the vector is compared to a set of allowed vectors from a table. As in scalar quantization, usually a mean squared error measure is used to select the closest vector from the quantization table. A weighted mean squared error measure is

often used to emphasize the components of the vector that are known to be perceptually more important.

Vector quantization is usually applied to the excitation signal and the LPC parameters. In the case of LPC coefficients, the range of the coefficients is unconstrained at least theoretically. This as well as stability problems due to slight errors in representation have resulted in first transforming the LPC coefficients to a more suitable parameter domain prior to quantization. The transformations allow the LPC coefficients to be represented with a set of parameters that have a known finite range and prevent instability or at least reduce its likelihood. Available methods include log-area ratios and inverse sine functions. A more computationally complex representation of the LPC coefficients is the line spectrum pair (LSP) representation. The LSPs provide a pseudo-frequency representation of the LPC coefficients and have been found to be capable of improving coding efficiency by more than other transformation techniques as well as having other desirable properties such as a simple way to guarantee stability of the LP synthesis filter.

Gain parameters and pitch periods are sometimes quantized this way. For instance, the GSM EFR coder quantizes the codebook gain differentially. A general technique for differential requantization will now be discussed.

Suppose G(n) is the parameter to be requantized and that the linear predictor used in the quantization scheme is denoted P(z) as shown in Figure 28. The quantized difference, R(n), is the actual coded domain parameter normally transmitted from the encoder to the decoder. This parameter is preferably intercepted by the network speech enhancement device and possibly modified to a new value, P(z). The operation of this method will now be explained with reference to Figure 28.

Suppose the speech enhancement algorithm required G(n) to be modified by an amount $\Delta G(n)$. The differential requantization scheme at the network device basically attempts to mimic the operation of the encoder. The basic idea behind this technique can be understood by first ignoring all the quantizers in the figure as well as the interconnections between the different systems. Then it is seen that the systems in the encoder and the network are both identical linear systems. The encoder has G(n) as its input while the network device has $\Delta G(n)$ as its input. Since they are preferably identical linear systems, it is realized that the two systems can be conceptualy combined to effectively result in a single system that has $(G(n)+\Delta G(n))$ as its input. Such a system preferably includes an output, $R_{new}(n)$, which is preferably be given by

$$R_{new}(n) = R(n) + \Delta R(n) \quad (17)$$

However, due to the quantizers which make these systems non-linear, this relationship is only approximate. Hence, the actual decoded parameter is preferably given by

$$G_{new}(n) = G(n) + \Delta G(n) + \text{quantization error} \quad (18)$$

where $\Delta G(n)$+ quantization error is the actual realized change in the parameter achieved by the network speech enhancement device.

The feedback of the quantity, $\Delta R(n)$, in the network requantization device can cause undesirable oscillatory effects if G(n) is not changing for long periods of time. This can have undesirable consequences to the speech signal especially if G(n) is a gain parameter. In the case of the GSM EFR codec, the G(n) corresponds to the

logarithm of the codebook gain. During silent periods, G(n) may remain at the same quantized level for long durations. During such silence, if attenuation of the signal is attempted by the network device by modifying G(n) by an appropriate amount $\Delta G(n)$, quasi-periodic modulation of the noise could occur resulting in a soft but disturbing buzz.

As an example, such oscillatory effects will be demonstrated when the GSM EFR codec is used. This linear predictor, P(z), preferably has four delay elements, each of which could take on one of 32 possible values. Hence the non-linear system in the ALC device can be in any one of over a million possible states at any given time. This is mentioned because the behavior of this non-linear system is heavily influenced by its initial conditions.

The reverberations in the actual realized gain, $G_{new}(n)$, will now be demonstrated in response to a step, $\Delta G(n)$, in the desired gain. For simplicity, it is assumed that the original transmitted parameter, R(n), is constant over 100 subframes, and that the memory of P(z) is initially zero. Figure 29(a) shows the step in the desired gain. Figure 29(b) shows the actual realized gain superimposed on the desired gain. Although the initial conditions and the value of G(n) will determine the exact behavior, the reverberations in the actual realized gain shown here are typical.

The reverberations can be eliminated by 'moving' the quantizer outside the feedback loop as shown in Figure 30. (In Figure 30, the computation of is unnecessary but is included for comparison to Figure 28.) Placing the quantizer outside the feedback loop results in the actual realized gain shown in Figure 29(c), superimposed on the desired gain. It should be noted that, although reverberations are eliminated, the average error (i.e. the average difference between the desired and actual realized

gains) is higher than that shown in Figure 29(b). Specifically, for this example, the average error during steady state operation of the requantizer with and without the quantizer in the feedback loop are 0.39dB and 1.03dB, respectively.

Hence a trade-off exists between accurate control of a differentially quantized parameter and potential oscillatory effects. However, through the use of a voice activity detector, it is possible to switch between the accurate scheme and the reverberation-free but less accurate scheme. The reverberation-free scheme would be used during silent periods while the more accurate scheme with the quantizer in the feedback loop would be used during speech. When switching between the schemes, the state of the predictor should be appropriately updated as well.

It should also be pointed out that the reverberation-free technique can be simplified as shown in Figure 31, resulting in some savings in computations. This is done by replacing the linear system 1/[1+P(z)] with the constant, 1/[1+P(1)]. This implementation is often found to be sufficient especially when the parameters are changed relatively infrequently. By infrequent changes, we mean that the average number of subframes between changes is much greater than the order of P(z).

Even when more sophisticated quantization schemes are used, the cost of re-encoding these parameters is still relatively small. With an understanding of how parameter modification can be practically effected even when the parameter is differentially quantized, the problems associated with coded domain noise reduction and echo suppression may be addressed.

A low complexity, low delay coded domain noise reduction method will now be discussed. The various coded domain parameters that could be used to effect noise reduction were discussed above. Of these parameters, it was determined that the two

gain parameters, the pitch gain, g_p, and the codebook gain, g_c, are most amenable to direct modification. Accordingly, the preferred embodiments will involve these parameters.

By way of example only, a commonly used subframe period of duration 5ms will be assumed. With the typical sampling rate of 8000Hz used in telephony applications, a subframe will consist of 40 samples. A sample index will be denoted using n, and the subframe index using ___. Since the coded parameters are updated at most once per subframe and apply to all the samples in the subframe, there will be no confusion if these coded parameters are simply indexed using m. Other variables that are updated or apply to an entire subframe will also be indexed in this manner. The individual samples within a subframe will be normally indexed using n. However, if more than one subframe is spanned by an equation, then it will make sense to index a sample, such as a speech sample, as s(n, m).

The speech synthesis model that is used in hybrid, parametric, time domain coding techniques can be thought of as time varying system with an overall transfer function, $H_m(z)$, at subframe m given by

$$H_m(z) = \frac{g_c(m)}{\left[1 - g_p(m)z^{-T(m)}\right]A_m(z)} \qquad (19)$$

with an excitation source provided by the fixed codebook (FCB). Another view that is closer to actual implementation is shown in Figure 32. The FCB output is indicated as C'(n). In Figure 32, the buffer of the long-term predictor (LTP) or pitch synthesis

filter is shown. Recall that the LTP has the transfer function $\frac{1}{(1-g_p z^{-T})}$, where both g_p and T are usually updated every subframe. According to this transfer function, the LP excitation would be computed for each subframe as

$$\begin{aligned} u(n) &= g_c(m)c^1(n) + g_p(m)b^1(n) \\ &= g_c(m)c^1(n) + g_p(m)u(n-T) \qquad (20) \\ n &= 0,1,...39 \end{aligned}$$

where b'(n) is obtained from the LTP buffer. The most recently computed subframe of LP excitation samples, u(n), are preferably shifted into the left end of the LTP buffer. These samples are also used to excite the LP synthesis filter to reconstruct the coded speech.

Using this viewpoint of the speech synthesis model, the two sources of the LP synthesis filter excitation, u(n), have been explicitly identified. These two excitation sources, denoted as b(n) and c(n), are called the pitch excitation and codebook excitation, respectively. Due to this two source viewpoint, the LTP is also often called the adaptive codebook, due to its ever-changing buffer contents, in contrast to the FCB. Obviously, the LTP output is not independent of the FCB output. Hence spectral subtraction concepts preferably are not directly applied to the two sources. However, it is noted that, due to the manner in which the encoder optimizes the coded domain parameters, the two sources have different characteristics. This difference in characteristic is taken advantage of to derive a noise reduction technique.

To achieve noise reduction, the gain factors, g_p and g_c that are received from the encoder are modified. This modification will be achieved by multiplying these

gain factors by the noise reduction gain factors, γ_p and γ_c, respectively, to generate an adjusted gain value. This will result in a modified time varying filter at the decoder given by

$$H_m(z) = \frac{\gamma_c(m)g_c(m)}{[1-\gamma_p(m)g_p(m)z^{-T(m)}]A_m(z)} \quad (21)$$

A preferred network noise reduction device is shown in Figure 33. In this embodiment, there are two decoders. A decoder 20 is termed the reference decoder and performs decoding of the coded speech received from the encoder, such as the speech encoder 10 shown in Figure 14. The decoding performed by decoder 20 may be complete or partial, depending on the particular codec. For the current embodiment, it is assumed that it performs complete decoding, producing the noisy speech output y(n). However, as described above, the embodiment also will operate with partial decoding. Essentially, decoding which does not substantially affect, for example, the power of the noise characteristic, can be avoided, thereby saving time.

The bottom half of Figure 33 shows a destination decoder 120. Using this decoder, the coded parameters may be optimized. This destination decoder mimics the actual decoder at the destination, such as the receiving handset. It produces the estimated clean speech output on a conductor 148. Note that, although drawn separately for clarity, some of the parts of the reference decoder and destination decoder model can be shared. For instance, the fixed codebook (FCB) signal is identical for both decoders.

Those skilled in communications will recognize that decoders 20 and 120 may be substituted for the following blocks of Figure 14:

Partial or Complete Decoding block;

Speech Level Measurement block;

Gain Determination block;

Multiply function having inputs SLRP and gain;

SLRP Requantization; and

Modify SLRP.

In addition, the Voice Activity function referred to in Figure 14 is incorporated into the Figure 33 embodiment. As a result, the speech decoder 12 shown in Figure 33 may be the same type of speech decoder shown in Figure 14.

More specifically, the Figure 33 decoders are useful in a communication system 8 using various compression code parameters, such as the parameters described in Figure 7, including codebook gain, pitch gain and codebook RPE pulses. Such parameters represent an audio signal having various audio characteristics, including a noise characteristic and signal to noise ratio (SNR). The Figure 33 apparatus provides an efficient technique for managing the noise characteristic. Decoders 20 and 120 may be implemented by a processor generally indicated by 150 which may include a noise reduction controller 160 which includes a VAD function. Processor 150 may comprise a microprocessor, a microcontroller or a digital signal processor, as well as other logic units capable of logical and arithmetic operations. Decoders 20 and 120 may be implemented by software, hardware or some combination of software and hardware.

Processor 150 responds to the compression code of the digital signals sent by encoder 10 on a network 11. Decoders 20 and 120 each read certain compression code parameters of the type described in Figure 7, such as codebook gain and pitch

gain. Processor 150 is responsive to the compression code to perform the partial decoding , if any, needed to measure the power of the noise characteristic. The decoding results in the decoded signals in the linear domain which simplify the task of measuring the noise power.

The reference decoder 20 receives the compression coded digital signals on terminals 13. Decoder 20 includes a fixed codebook (FCB) function 22 which generates codebook vectors C'(n) that are multiplied or scaled by codebook gain g_c in a multiply function 24. The codebook gain is read by processor 150 from the compressed code signals received at terminals 13. The multiply function generates scaled codebook vectors c(n) which are supplied to a pitch synthesis filter 26. Processor 150 calculates the power P_c of the scaled codebook vectors as shown in equation 31. The power is used to adjust the pitch gain. Processor 150 reduces the codebook gain to attenuate the scaled codebook vector contribution to the noise characteristic.

Filter 26 includes a long term predictor (LTP) buffer 28 responsive to the scaled codebook vectors c(n) to generate sample vectors. The samples are scaled by the pitch gain g_p in a multiply function 30 to generate scaled samples b_{ref} (n) that are processed by an adder function 32. Processor 150 increases the pitch gain to increase the contribution of the scaled samples in order to manage the noise characteristic as indicated in equations 30-33. Processor 150 determines the power of the scaled samples P_{bref}. A similar power P_b is generated by decoder 120. The two powers are used to adjust the pitch gain as indicated by equations 30 and 33.

Filter 26 generates a total codebook excitation vector or LPC excitation vector u(n) at its output. Processor calculates the power P_u of vector u(n) and uses the power to adjust the pitch gain as indicated in equation 32.

The vector u(n) excites an LPC synthesis filter 34 like the one shown in Figure 6. The output of filter 34 is returned to controller 160.

Decoder 120 includes many functions which are identical to the functions described in connection with decoder 20. The like functions bear numbers which are indexed by 100. For example, codebook 22 is identical to codebook 122. Decoder 120 includes multiplier functions 140 and 142 which are not included in decoder 20. Multiplier function 140 receives γ_p as an input which is defined in equation 33. As shown in equation 30, the value of γ_p depends in part on a ratio of powers previously described. Multipler function 142 receives γ_c as an input which is defined in equation 28. As a result of multiplier functions 140 and 142, decoder 120 uses a pitch synthesis filter 144 which is different from pitch synthesis filter 26.

As explained by the equations in general and equations 21-33 in particular, processor adjusts the codebook gain and/or pitch gain to manage the noise characteristic of the signals received at terminals 13. The adjusted gain values are quantized in the manner previously described and the quantized parameters are transmitted on an output network 15 through a terminal 16.

The basic single-band noise suppressor discussed above can be implemented in the coded domain. Since $g_c(m)$ is the DC gain of the time-varying filter given in equation (19), this DC gain can be modified by setting $\gamma_c(m)$ as

$$\gamma_c(m) = \max\left(1 - \frac{P_w(m)}{P_y(m)}, E\right) \quad (22)$$

where $P_w(m)$ and $P_y(m)$ are the noise power and total power estimate, respectively, at subframe m, respectively. Also, E is the maximum loss that can be applied by the single-band noise suppressor. It is usually set to a small value such as 0.1. Such a DC gain control system will suffer from severe noise modulation because the noise power fluctuates in sync with the speech signal. This can be perceptually annoying and one way to compensate for this is by trading off the amount of noise suppression for the amount of noise modulation.

A coded domain noise reduction method may be derived that is superior to the that in equation (20). The two parameters, γ_p and γ_c, can be controlled in the time-varying system $H_m(z)$. Due to the recursive nature of the decoder, the joint optimization of both gain factors to achieve noise reduction is rather complex. This is because the modification of the present value of ___ would have implications on future values of g_p. Hence such optimization would preferably determine $\gamma_c(m)$ and $\gamma_p(m+l)$ where l depends on the time-varying pitch period, T(m). Even a sub-optimal optimization would require knowledge of coded parameters at least a few subframes into the future. This would require crossing frame boundaries and has severe practical implications. First, more buffering would be required. More

importantly, additional delay would be incurred which may be unacceptable especially in cellular and packet networks. Thus, the problem is preferably approached in a manner that does not require knowledge of future frames.

The basic idea behind the technique will first be stated. During silence as indicated by a voice activity detector, it is safe to perform the maximum attenuation on the signal by limiting the DC gain of $H_m(z)$ by controlling γ_c. At the beginning and trailing ends of speech, the γ_c will be allowed to rise and fall appropriately. However, during voiced speech, the LTP excitation output contributes to a large amount of the resulting signal power and has a better SNR relative to the FCB excitation output. Hence, during voiced speech, we can also perform a limited amount of attenuation of the FCB output. To compensate for the eventual loss of power in the noise-reduced decoded speech signal, γ_p will be carefully boosted. γ_p and γ_c will be optimized in two stages.

First, the optimization of γ_c will be considered. To reduce the noise effectively, γ_c should preferably be driven close to zero or some maximum loss, $0\langle E\langle 1$. The trade-off with using a high loss is that the decoded speech signal would also be attenuated. To reflect this tug-of- war between maintaining the decoded speech level which requires that $\gamma_c = 1$ and obtaining effective noise reduction which requires that $\gamma_c = E$ can be stated in terms of a cost function, F, as follows:

$$F(\gamma_c, \lambda_1, \lambda_2) = \lambda_1(\gamma_c - E)^2 + \lambda_2(\gamma_c - 1)^2 \quad (23)$$

Here λ_1 and λ_2 are suitable weights to be determined. By minimizing this cost function, an optimal amount of DC gain reduction may be achieved. In this context, one set of suitable weights that have proven to provide consistently good results will be considered. Nevertheless, other suitable weights may be formulated that perform just as well.

During silence, we would like to achieve the maximum amount of noise suppression. Hence λ_1 should preferably be large during silence gaps and small during speech. A suitable continuous measure that can achieve such a weighting is the SNR measured using the reference decoder, denoted as SNR_{ref}. The first weight may be set as

$$\lambda_1 = \frac{1}{SNR_{ref}(m)} \tag{24}$$

A voice activity detector can be used to demarcate the silence segments from the speech segments in the reference decoder's output signal, y(n). The background noise power, P_w, can be estimated during silence gaps in the decoded speech signal y(n). The recursive averager of equation (11) with a pole at 15999/16000 and unity DC gain is found to be a suitable means for updating the background noise power during such silence gaps. This large time constant is suitable since noise can be assumed to be relatively stationary. The power, P_y, of the signal, y(n), can also be measured using a similar recursive average or other means. If a recursive average is utilized, an averager with a pole at 127/128 and unity DC gain was found to be suitable. Then, SNR_{ref} can be estimated as

$$SNR_{ref} = \max\left[O, \frac{P_y - P_w}{P_w}\right], P_w \rangle O$$

(25)

Here, the maximum function disallows meaningless negative values for the SNR_{ref} that may occur. It is assumed that the noise power estimation algorithm always ensures that P_w is greater than zero.

If only λ_1 was used and λ_2 was set to unity, then the γ_c will rise and fall with the SNR_{ref}. However, during voiced speech which typically also has higher SNR, γ_c is preferably attenuated to some extent. This would reduce the overall amount of noise during voiced speech as the FCB models the majority of the noise signal during voiced speech. Hence the noise modulation that typically occurs in single-band noise reduction systems will be reduced. An appropriate parameter that reflects the presence of voiced speech is necessary. The ratio, $P_{b,ref} / P_{c,ref}$, where P_b and P_c are the short-time powers of the reference decoder signals, $b_{ref}(n)$ and $C_{ref}(n)$, indicated in Figure 33, reflect the presence of voiced speech. Alternatively, the pitch gain, $g_p(m)$, which also reflects the amount of correlation in the speech, may be used. Recall that the pitch gain is the result of an optimization procedure at the encoder that determines the pitch synthesis filter. In essence, this procedure finds a past sequence from the LTP buffer that has the best correlation with the sequence to be embodied. Therefore, if the correlation is high, then the pitch gain would also be correspondingly high. As such, the remaining weight may be specified to be inversely proportional to the pitch gain:

$$\lambda_2 = \lambda \frac{1}{g_p(m)}$$

(26)

By specifying λ_2 in this manner, keeping γ_c close to one during voiced speech is deemphasized.

The parameter λ is preferably empirically determined. It is quite common to have parameters that require to be tuned based on perceptual tests in speech enhancement algorithms.

Thus, the resulting cost function to be minimized is

$$F(\gamma_c, \lambda) = \frac{1}{SNR_{ref}}(\gamma_c - E)^2 + \lambda \frac{1}{g_p}(\gamma_c - 1)^2$$

(27)

By taking the derivative of F with respect to γ_c and setting it to zero, the optimum value of γ_c is determined to be

$$\gamma_c = \frac{E + \lambda \frac{SNR_{ref}}{g_p}}{1 + \lambda \frac{SNR_{ref}}{g_p}}$$

(28)

where λ will be optimized empirically. Now γ_c still generally rises and falls in sync with the SNR_{ref}. However, a smaller γ_c may result even if SNR_{ref} is large if, in addition, g_p is also large.

By determining γ_c according to equation (28), the overall signal power of the clean speech estimate, $\hat{S}(n)$, may be reduced. This power loss can be compensated to

some extent by increasing γ_p appropriately. First, the characteristics of the LTP or pitch synthesis filter used in the coder will be considered.

The pitch synthesis filter is basically a comb filter. The first 1kHz range of the magnitude frequency response of comb filters obtained when the pitch period of T = 40 is shown in Figure 34. Two curves are shown, one corresponding to a pitch gain of 0.1 and the other 0.9. We note that since only the pitch gain and pitch period are used to specify the pitch synthesis filter, there is no DC gain factor available to simultaneously control the amount of gain at both the spectral peaks and the valleys. Another point to note is that some encoders allow pitch gains greater than one. Theoretically, this results will result in an unstable comb filter. However, due to the manner in which the optimization procedure attempts to match the synthetic signal to the original speech signal, no actual instability results. Another way to look at this is to think of the FCB output as being designed in such a manner that never actually results in instability.

By multiplying γ_p with the original pitch gain, g_p, it is possible to cause instability or at least large undesirable fluctuations in power. It is noted that the increase, I_{peak}, in the magnitude frequency response at a spectral peak of the comb filter due to applying γ_p is given by

$$I_{peak} = 20 \log_{10}\left(\frac{1-g_p}{1-\gamma_p g_p}\right), \gamma_p g_p < | \text{ and } g_p < |$$

(29)

Typical values of I_{peak} are illustrated in Figure 35 for two values of g_p that are common during voiced speech in a noisy speech signal. From this figure, it is

seen that large gains can be induced at the spectral peaks. It should be noted that the spectral valleys are also attenuated.

Some level of noise reduction is achieved by the attenuation of the spectral valleys. However, at the same time, the noise present in the spectral peaks of the LTP gets amplified. Overall, this can result in the noise being shaped to have a harmonic character. Such harmonically shaped noise, if present in significant amounts, can make the speaker's voice sound somewhat nasal in character. Thus, great care should be taken when boosting γ_p. Amplification to compensate for power loss may be performed only if $g_p \langle |$ and the amplified pitch gain should satisfy $\gamma_p g_p \langle |$.

Preferably, one could compensate for the power loss in the LTP excitation output. To achieve this power compensation accurately, a first possibility for γ_p would be computed as

$$\gamma_{p,1} = \frac{P_{b,ref}}{P_b} \tag{30}$$

This could sometimes result in instability in total LP excitation. To compensate for power loss and ensure stability, $\gamma_{p,1}$ could be compared with $\gamma_{p,2}$ computed as $\gamma_{p,2} = \sqrt{P_{u,ref}/P_u}$. However, this involves a trial and error process as P_u depends on γ_p. An alternative is to approximate P_u as $P_u = \gamma_c^2 P_c + \gamma_p^2 P_b$. Then, the stability condition can be specified as

$$\gamma_c^2 P_c + \gamma_p^2 P_b \le P_{u,ref} \tag{31}$$

which would give the second possible value for γ_p as

$$\gamma_{p,2} = \sqrt{\frac{P_{u,ref} - \gamma_c^2 P_c}{P_b}}$$

(32)

Then, γ_p should be determined as the minimum of the two quantities in equations (30) and (32). A further check to ensure that the resulting filter will be stable may be performed. In this case, γ_p is preferably chosen as

$$\gamma_p = \begin{cases} \min[\gamma_{p,1}, \gamma_{p,2}] & \text{if } \min[\gamma_{p,1}, \gamma_{p,2}] \text{ gp} \langle \\ 1 & \text{otherwise} \end{cases}$$

(33)

However, as the risk of instability is small, this last check may be avoided. Furthermore, the criterion in equation (32) ensures that the resulting LTP output will be stable.

Two additional embodiments for coded domain noise reduction (CDNR) will be discussed in connection with Figure 36. In one of the two embodiments, only the codebook gain parameter (g_c) is modified. In the second embodiment, both the codebook gain and pitch gain (g_p) are modified. The first embodiment is suitable for low levels of noise while the second embodiment is suitable for higher noise conditions.

CDNR by Codebook Gain Attenuation

Figure 36 shows a novel implementation of CDNR. Given the coded speech parameters corresponding to each frame of speech, the uncoded speech is

reconstructed using the appropriate decoder. A silence detector (also referred to as a voice activity detector) is used to determine whether the frame corresponds to speech or silence. If the frame is silence, then the background noise power is estimated. At all times, the total power of the signal is estimated. Using the total power and noise power, it is possible to infer the relative amount of noise in the signal, such as by computing the signal-to-noise ratio. Based on these power estimates, the dequantized codebook gain parameter is attenuated, and then quantized again. This new quantized codebook gain parameter substitutes the original one in the bit-stream.

The careful attenuation of the codebook gain parameter can result in noise reduction in the case of noisy coded speech. Many attenuation methodologies can be formulated. Before describing any methods, the notation used is first described.

We assume that the noisy uncoded speech, $y(n)$, is given by

$$y(n) = s(n) + w(n) \qquad (34)$$

where $s(n)$ is the clean uncoded speech and $w(n)$ is the additive noise. The power estimates, $P_y(n)$ and $P_w(n)$, are the noisy uncoded speech power and the noise power, respectively. In Figure 36, $P_y(n)$ is measured in the block labeled "Total power estimator" and $P_w(n)$ is measured in the block labeled "Noise power estimator". Power estimates may be performed in a variety of ways. One example approach is the recursive formula given by $P_y(n) = \beta P_y(n) + (1-\beta)\|y(n)\|$, with $\beta = 0.992$, and a similar formula for the noise is given by $P_w(n) = \beta P_w(n) + (1-\beta)\|w(n)\|$ with $\beta = 0.99975$.

The codebook gain factor, g_c, is multiplied by an attenuation factor, γ_c, to obtain an attenuated codebook gain factor. Two alternative methods for determining γ_c are as follows:

$$\gamma_c = \max\left[0{,}1 - \mu\frac{P_w}{P_y}\right] \quad (35)$$

$$\gamma_c = \min\left[1{,}0.2 + \mu\frac{P_y - P_w}{P_w}\right] \quad (36)$$

In most vocoders, the codebook gain parameters are defined every subframe. If this is the case, the formulae are evaluated using the power estimates computed during the last sample of the corresponding subframe. In both the above approaches, the attenuation factor depends on the signal-to-noise ratio of the uncoded speech. In formula (35), a suitable value for μ are in the range from 1 to 1.5. In formula (36), a suitable value for μ is 0.8.

CDNR by Optimization of Gain Factors

Partial Decoding

The decoding of signals may be complete or partial depending on the vocoder being used for the encode and decode operations. Some examples of situations where partial decoding suffices are listed below:

In code-excited linear prediction (CELP) vocoders, a post-filtering process is performed on the signal decoded using the LPC-based model. This post-filtering

process reduces quantization noise. However, since it does not significantly affect the power estimates, the post-filtering stage can be avoided for economy.

Under TFO in GSM networks, the CDNR device may be placed between the base station and the switch (known as the A-interface) or between the two switches. Since the 6 MSBs of each 8-bit sample of the speech signal corresponds to the PCM code as shown in Figure 3, it is possible to avoid decoding the coded speech altogether in this situation. A simple table-lookup is sufficient to convert the 8-bit companded samples to 13-bit linear speech samples using A-law companding tables. This provides an economical way to obtain a version of the speech signal without invoking the appropriate decoder. Note that the speech signal obtained in this manner is somewhat noisy, but has been found to be adequate for the measurement of the power estimates.

Coded Parameter Modification

Minimal Delay Technique

Large buffering, processing and transmission delays are already present in cellular networks without any network voice quality enhancement processing. Further network processing of the coded speech for speech enhancement purposes will add additional delay. Minimizing this delay is important to speech quality. In this section, a novel approach for minimizing the delay is discussed. The example used is the GSM FR vocoder.

Figure 7 shows the order in which the coded parameters from the GSM FR encoder are received. A straightforward approach involves buffering up the entire 260 bits for each frame and then processing these buffered bits for coded domain echo control purposes. However, this introduces a buffering delay of about 20ms plus the processing delay.

It is possible to minimize the buffering delay as follows. First, note that the entire first subframe can be decoded as soon as bit 92 is received. Hence the first subframe may be processed after about 7.1ms (20ms times 92/260) of buffering delay. Hence the buffering delay is reduced by almost 13ms.

When using this novel low delay approach, the coded LPC synthesis filter parameters are modified based on information available at the end of the first subframe of the frame. In other words, the entire frame is affected by the echo likelihood computed based on the first subframe. In experiments conducted, no noticeable artifacts were found due to this 'early' decision.

Update of Error Correction/Detection Bits and Framing Bits

When applying the novel coded domain processing techiques described in this report for removing or reducing noise, some are all of the bits corresponding to the coded parameters are modified in the bit-stream. This may affect other error-correction or detection bits that may also be embedded in the bit-stream. For instance, a speech encoder may embed some checksums in the bit-stream for the decoder to verify to ensure that an error-free frame is received. Such checksums as well as any

parity check bits, error correction or detection bits, and framing bits are updated in accordance with the appropriate standard, if necessary.

Figure 38shows a technique for coded domain noise reduction by modification of the codebook vector parameter. In the preferred mode, noise reduction is performed in two stages. The first stage involves modification of the codebook gain as discussed earlier.

In the second stage, the codebook vector is optimized to minimize the noise. In essence, for each subframe, several codebook vector patterns are attempted that vary from the original received codebook vector. For each codebook vector pattern, the partial decoding is performed and the noise power is estimated. The best codebook vector pattern is determined as the one that minimizes the noise power. In practice, a fixed number of iterations or trials are performed.

For example, in the GSM FR vocoder (Reference [1]), the codebook vector pattern for each subframe has 40 positions, of which 13 contain non-zero pulses. In our preferred mode, the positions of the 13 non-zero pulses are not modified. Only their amplitudes are varied in each trial. The non-zero pulses are denoted by $x_M(i), i = 0,1,...,12$. Note that each pulse may be one of the following amplitude values only: $\pm 28672, \pm 20480, \pm 12288, \pm 4096$. The codevector optimization is described by the following steps:

Using the original codebook vector, modified codebook gain parameter, and the remainder of the original parameters, partially decode the signal.

Estimate the noise power in the decoded signal and save this value.

Set $i=0, j=1$.

In the original codebook vector, modify the i^{th} pulse $x_M(i)$ to be j levels of amplitude smaller but of the same sign, so as to obtain a modified codebook vector. If already at the lowest level for the given sign, then change the sign.

Using the modified codebook vector, modified codebook gain parameter, and the remainder of the original parameters, partially decode the signal.

Estimate the noise power in the decoded signal and save this value.

Repeat steps 2 to 4 for $i=1,2,...,12$.

Set $i=0, j=2$ and repeat steps 2 to 5 for this new value of j.

At this point, the partial decoding would have been performed 27 times. Pick the codebook vector that resulted in the minimum amount of noise.

It is straightforward to modify the above search technique for the codebook vector optimization, or implement other codebook vector search techniques such as those used in codebook-excited linear prediction (CELP) vocoders.

CDNR by modification of the representation of the LPC parameters

A commonly used technique for the representation of the LPC parameters is considered as an example. This representation, called the line spectral pairs (LSPs) or frequencies (LSFs) has become widely used in many vocoders, e.g. the GSM EFR,

due to its good properties in terms of quantization and stability, as well as interpretation. The LSFs are a pseudo-frequency representation of the LPC parameters. This allows the quantization techniques to incorporate information about the spectral features that are known to be perceptually important. Another advantage of LSFs is that they facilitate smooth frame-to-frame interpolation of the LPC synthesis filter.

As another example, LPC parameters also are represented by log area ratios in the GSM FR vocoder.

LSFs may be directly modified for speech enhancement purposes. A technique that directly adapts the LSFs to attain a desired frequency response for use in a coded domain noise reduction system is described in the following. This general technique may be applied to modify the LSFs, for example, received from a GSM EFR encoder.

In a coded domain noise reduction technique, the adaptive technique may be used to alter the spectral shape of the LPC synthesis filter, $1/A(z) = 1/\left[1 - \sum_{i=1}^{p} a_i z^{-i}\right]$, when represented in terms of LSFs, to attain a desired spectrum according to spectral subtraction principles.

If the denominator polynomial, $A(z) = \sum_{i=1}^{p} a_i z^{-i}$, of the LPC synthesis filter transfer function has p coefficients, then an anti-symmetric and a symmetric polynomial can be derived as follows:

$$P(z) = A(z) - z^{-(p+1)}A(z^{-1})$$
$$Q(z) = A(z) + z^{-(p+1)}A(z^{-1})$$

Note that $A(z)$ can be recovered as $A(z) = \frac{1}{2}[P(z) + Q(z)]$.

The roots of these auxiliary polynomials are the LSPs and their angular frequencies are called the LSFs. Basically, each polynomial can be thought of as the transfer functions of a $(p+1)$th order predictor derived from a lattice structure. The first p stages of each of these predictors have the same response as the $A(z)$. $P(z)$ and $Q(z)$ have an additional stage each with reflection coefficients -1 and +1, respectively.

These auxiliary polynomials have some interesting properties. Given that $A(z)$ is minimum phase, the two important properties of $P(z)$ and $Q(z)$ can be proven. First, all the zeros of both these polynomials are on the unit circle. Second, the zeros of $P(z)$ and $Q(z)$ are interlaced. Furthermore, if the zeros remain interlaced through a quantization process, then the resulting $A(z)$ obtained is guaranteed to be minimum phase.

In addition to these useful properties, the LSFs have a pseudo-frequency interpretation that is often useful in the design of quantization techniques. Figure 39 shows a randomly generated set of LSFs and the frequency response of the corresponding linear predictor which has 10 coefficients. The solid vertical lines are the angles of the roots of $P(z)$ while the dashed lines are the angles of the roots of $Q(z)$. Note that the angles completely specify the roots of these polynomials which all lie on the unit circle.

A loose spectral interpretation of the LSFs comes about from the observation that the sharp valleys tend to be bracketed by the LSFs. Thus, the sharp peaks of each formant region of the LPC synthesis filter, $1/A(z)$, which are perceptually important in speech, tend to correspond to a pair of closely spaced LSFs.

We now derive a novel technique for the direct adaptation of the LSFs to achieve a desired spectral response. We constrain our discussion to even orders of p only. This is not a major restriction as speech coders usually use even ordered $A(z)$ functions. Use of an odd number of coefficients in $A(z)$ would be somewhat of a waste since DC components are usually removed prior to speech processing and coding.

First, the polynomials, $P(z)$ and $Q(z)$ are factorized as

$$P(z) = (1 - z^{-1}) \prod_{i=1}^{p/2} \left(1 + c_i z^{-1} + z^{-2}\right)$$
$$Q(z) = (1 + z^{-1}) \prod_{i=1}^{p/2} \left(1 + d_i z^{-1} + z^{-2}\right)$$

where $c_i = -2\cos\theta_{c,i}$ and $d_i = -2\cos\theta_{d,i}$. The $\{\theta_{c,i}, \theta_{d,i}\}$ are the LSFs specified in radians. The $\{c_i, d_i\}$ are termed the LSFs in the cosine domain. Note that if $A(z)$ is minimum phase, then

$$0 \le \theta_{c,1} < \theta_{d,1} < \theta_{c,2} < \theta_{d,2} < \ldots < \theta_{c,p/2} < \theta_{d,p/2} \le \pi$$

will be true if the LSFs are sorted and labelled appropriately.

The power or magnitude squared frequency response of $A(z)$ is

$$|A(\omega)|^2 = 0.25|P(\omega)|^2 + 0.25|Q(\omega)|^2$$

where it can be shown that $|P(\omega)|^2$ and $|Q(\omega)|^2$ are given by

$$|P(\omega)|^2 = 2(1-\cos\omega)\prod_{i=1}^{p/2}\left[c_i^2 + 4c_i\cos\omega + (2+2\cos 2\omega)\right]$$
$$|Q(\omega)|^2 = 2(1+\cos\omega)\prod_{i=1}^{p/2}\left[c_i^2 + 4c_i\cos\omega + (2+2\cos 2\omega)\right]$$

Next, we utilize the method of steepest descent to adapt the LSFs in the cosine domain, $\{c_i, d_i\}$, to achieve the power frequency response specified at a set of frequencies $\{\omega_k\}$. Suppose the specified power frequency response is given as $\{A_k^2\}$ at N different frequencies. Then we write the squared error between $\{A_k^2\}$ and the actual power frequency response $\{|A(\omega_k)|^2\}$ of $A(z)$ at frequencies $\{\omega_k\}$ as a function of the $\{c_i, d_i\}$. This error function is

$$F(\{c_i, d_i\}) = \sum_{k=0}^{N-1}\left[A_k^2 - |A(\omega_k)|^2\right]$$
$$= \sum_{k=0}^{N-1}\left[A_k^2 - 0.25|P(\omega_k)|^2 - 0.25|Q(\omega_k)|^2\right]$$

According to the method of steepest descent, we can update the LSFs in the cosine domain at the $(n+1)$ th iteration in terms of the values at the n th iteration as follows:

$$c_i(n+1) = c_i(n) - \mu\frac{\partial F}{\partial c_j}$$
$$d_i(n+1) = d_i(n) - \mu\frac{\partial F}{\partial d_j}$$

where μ is an appropriate step-size parameter.

In our preferred mode, the value of μ is set to 0.00002.

We have described a method for directly modifying the coded parameters, particularly the line spectral frequencies which are a representation of the LPC parameters. Using this method, the frequency response of the LPC synthesis filter can be modified to have a desired frequency response. For noise reduction purposes, the desired frequency response of the LPC synthesis filter can be computed based on, for example, standard noise reduction techniques such as spectral subtraction. In summary, the compression code parameters are modified to reduce the effects of noise. More specifically, the LPC coefficients or one of their representations (e.g., line spectral frequencies or log-arc ratios) are modified to atenuate the noise in spectral regions affected by noise.

Those skilled in the art of communications will recognize that the preferred embodiments can be modified and altered without departing from the true spirit and scope of the invention as defined in the appended claims. For example, the ALC techniques described in the specification also apply to NR techniques.

What is claimed is:

1. In a communication system for transmitting digital signals using a compression code comprising a predetermined plurality of parameters including a first parameter, said parameters representing an audio signal, said audio signal having a plurality of audio characteristics including a noise characteristic, said compression code being decodable by a plurality of decoding steps, apparatus for managing the noise characteristic comprising:

a processor responsive to said compression code of said digital signals to read at least said first parameter, and responsive to said compression code and said first parameter to generate an adjusted first parameter and to replace said first parameter with said adjusted first parameter.

2. Apparatus, as claimed in claim 1, wherein said processor performs said plurality of decoding steps by performing first decoding steps to generate first decoder signals resulting in a noisy speech signal and second decoding steps to generate second decoder signals resulting in an estimated clean speech signal, and wherein said processor responds at least to said first decoder signals and said second decoder signals and said first parameter to generate said adjusted first parameter.

3. Apparatus, as claimed in claim 1, wherein said first parameter comprises codebook gain, and wherein said processor modifies said codebook gain to modify the codebook vector contribution to said noise characteristic.

4. Apparatus, as claimed in claim 1, wherein said first parameter comprises codebook gain, wherein said plurality of parameters further comprises pitch gain, wherein said plurality of characteristics further comprises signal to noise ratio and

wherein said processor is responsive to said codebook gain, said pitch gain and said signal to noise ratio to generate said adjusted first parameter, and wherein said adjusted first parameter comprises an adjusted codebook gain.

5. Apparatus, as claimed in claim 4, wherein said signal to noise ratio comprises a ratio involving noisy signal power and noise power of said audio signal.

6. Apparatus, as claimed in claim 1, wherein said first parameter comprises pitch gain, wherein said plurality of parameters further comprise codebook gain, wherein said processor performs said plurality of decoding steps by generating a codebook vector, wherein said processor scales said codebook vector by said codebook gain to generate a scaled codebook vector, wherein said processor comprises at least a first buffer responsive to said scaled codebook vector to generate first samples based on pitch period, wherein said processor scales said first samples by said pitch gain to generate first scaled samples, and wherein said processor modifies said pitch gain to modify the contribution of said first scaled samples in order to manage said noise characteristic.

7. Apparatus, as claimed in claim 1, wherein said first parameter comprises pitch gain, wherein said plurality of characteristics further comprises signal to noise ratio, wherein said processor is responsive to said pitch gain and said signal to noise ratio to generate said adjusted first parameter, and wherein said adjusted first parameter comprises an adjusted pitch gain.

8. Apparatus, as claimed in claim 7, wherein said signal to noise ratio comprises a ratio involving noisy signal power and noise power of said audio signal.

9. Apparatus, as claimed in claim 1, wherein said first parameter comprises pitch gain, wherein said plurality of parameters further comprise codebook gain, wherein said processor performs said plurality of decoding steps to generate a codebook vector,

wherein said processor scales said codebook vector by said codebook gain to generate a scaled codebook vector, wherein said processor generates a power signal representing the power of said scaled codebook vector, wherein said processor is responsive to said pitch gain and said power signal to generate said adjusted first parameter, and wherein said adjusted first parameter comprises an adjusted pitch gain.

10. Apparatus, as claimed in claim 1, wherein said first parameter comprises pitch gain, wherein said processor comprises at least a first buffer generating at least first samples based on pitch period, wherein said processor scales said first samples by said pitch gain to generate at least first scaled samples, wherein said processor generates at least a first power signal representing the power of said first scaled samples, and wherein said processor is responsive at least to said pitch gain and said first power signal to generate said adjusted first parameter, and wherein said adjusted first parameter comprises an adjusted pitch gain.

11. Apparatus, as claimed in claim 10, wherein said processor comprises a second buffer responsive in part to said first power signal to generate second samples based on pitch period, wherein said processor scales said second samples by said pitch gain to generate second scaled samples, wherein said processor generates a second power signal representing the power of said second scaled samples and wherein said processor is responsive to said pitch gain, said first power signal and said second power signal to generate said adjusted first parameter.

12. Apparatus, as claimed in claim 11, wherein said first buffer and said second buffer each comprises a long-term predictor buffer.

13. Apparatus, as claimed in claim 1, wherein said first parameter comprises pitch gain, wherein said plurality of parameters further comprises a codebook gain,

wherein said processor comprises a pitch synthesis filter, wherein said processor performs said plurality of decoding steps to generate a first vector, wherein said processor scales said first vector by said codebook gain to generate a scaled codebook vector, wherein said processor filters said scaled codebook vector through said pitch synthesis filter to generate a second vector, wherein said processor generates a power signal representing the power of said second vector, wherein said processor is responsive to said pitch gain and said power signal to generate said adjusted first parameter, and wherein said adjusted first parameter comprises an adjusted pitch gain.

14. Apparatus, as claimed in claim 13, wherein said first vector comprises a codebook excitation vector and wherein said second vector comprises an LPC excitation vector.

15. Apparatus, as claimed in claim 1, wherein said first parameter comprises a codebook vector comprising pulses using variable sets of amplitudes, wherein said processor analyzes said sets to identify the powers of said noise characteristic represented by said sets, wherein said processor identifies a first set representing a power less than the power represented by said sets other than said first set, and wherein said processor adjusts said pulses according to said first set to generate said adjusted parameter.

16. Apparatus, as claimed in claim 1, wherein said plurality of decoding steps further comprises at least one decoding step that does not substantially affect the management of the noise characteristic and wherein said processor avoids performing said at least one decoding step.

17. Apparatus, as claimed in claim 16, wherein said at least one decoding step comprises post-filtering.

18. Apparatus, as claimed in claim 1, wherein said compression code comprises a linear predictive code.

19. Apparatus, as claimed in -claim 1, wherein said compression code comprises regular pulse excitation – long term prediction code.

20. Apparatus, as claimed in claim 1, wherein said compression code comprises code-excited linear prediction code.

21. Apparatus, as claimed in claim 1, wherein said first parameter is a quantized first parameter and wherein said processor generates said adjusted first parameter in part by quantizing said adjusted first parameter before replacing said first parameter with said adjusted first parameter.

22. Apparatus, as claimed in claim 1, wherein said compression code is arranged in frames of said digital signals and wherein said frames comprise a plurality of subframes each comprising said first parameter, wherein said processor is responsive to said compression code to read at least said first parameter from each of said plurality of subframes, and wherein said processor replaces said first parameter with said adjusted first parameter in each of said plurality of subframes.

23. Apparatus, as claimed in claim 22, wherein said processor replaces said first parameter with said adjusted first parameter for a first subframe before processing a subframe following the first subframe to achieve lower delay.

24. Apparatus, as claimed in claim 1, wherein said compression code is arranged in frames of said digital signals and wherein said frames comprise a plurality of subframes each comprising said first parameter, wherein said processor begins to perform said decoding steps during a first of said subframes to generate a plurality of said decoded signals, reads said first parameter from a second of said subframes

occurring subsequent to said first subframe, generates said adjusted first parameter in response to said decoded signals and said first parameter, and replaces said first parameter of said second subframe with said adjusted first parameter.

25. Apparatus, as claimed in claim 1, wherein said processor is responsive to said compression code to perform at least one of a plurality of said decoding steps to generate decoded signals and wherein said processor is responsive to said decoded signals and said first parameter to generate said adjusted first parameter.

26. Apparatus, as claimed in claim 1, wherein said first parameter is selected from the group consisting of codebook vector, codebook gain, pitch gain and LPC coefficients representations, including line spectral frequencies and log area ratios.

27. Apparatus, as claimed in claim 1, wherein said audio signals have spectral regions affected by said noise characteristic, wherein said first parameter comprises a representation of LPC coefficients, wherein said processor is responsive to said compression code and said representation to determine said spectral regions affected by noise and to generate said adjusted first parameter to manage said noise characteristic in said regions, and wherein said adjusted first parameter comprises an adjusted representation of LPC coefficients.

28. Apparatus, as claimed in claim 27, wherein said representation of LPC coefficients is selected from the group consisting of line spectral frequencies and log area ratios.

29. In a communication system for transmitting digital signals comprising code samples, said code samples comprising first bits using a compression code and second bits using a linear code, said code samples representing an audio signal, said audio signal

having a plurality of audio characteristics including a noise characteristic, apparatus for managing the noise characteristic without decoding said compression code comprising:

a processor responsive to said second bits to adjust said first bits and said second bits, whereby the noise characteristic in the digital signals is controlled.

30. Apparatus, as claimed in claim 29, wherein said linear code comprises pulse code modulation (PCM) code.

31. Apparatus, as claimed in claim 29, wherein said compression code samples conform to the tandem-free operation of the global system for mobile communications standard.

32. Apparatus, as claimed in claim 29, wherein said first bits comprise the two least significant bits of said samples and wherein said second bits comprise the 6 most significant bits of said samples.

33. Apparatus, as claimed in claim 32, wherein said 6 most significant bits comprise PCM code.

34. In a communication system for transmitting digital signals using a compression code comprising a predetermined plurality of parameters including a first parameter, said parameters representing an audio signal, said audio signal having a plurality of audio characteristics including a noise characteristic, said compression code being decodable by a plurality of decoding steps, a method of managing the noise characteristic comprising:

reading at least said first parameter;

generating an adjusted first parameter in response to said compression code and said first parameter; and

replacing said first parameter with said adjusted first parameter.

35. A method, as claimed in claim 34, and further comprising:

performing said plurality of decoding steps by performing first decoding steps to generate first decoder signals resulting in a noisy speech signal and second decoding steps to generate second decoder signals resulting in an estimated clean speech signal; and

responding at least to said first decoder signals and said second decoder signals and said first parameter to generate said adjusted first parameter.

36. A method, as claimed in claim 34, wherein said first parameter comprises codebook gain, and wherein said method further comprises modifying said codebook gain to modify the codebook vector contribution to said noise characteristic.

37. A method, as claimed in claim 34, wherein said first parameter comprises codebook gain, wherein said plurality of parameters further comprises pitch gain, wherein said plurality of characteristics further comprises signal to noise ratio and wherein said generating comprises generating said adjusted first parameter in response to said codebook gain, said pitch gain and said signal to noise ratio, and wherein said adjusted first parameter comprises an adjusted codebook gain.

38. A method, as claimed in claim 37, wherein said signal to noise ratio comprises a ratio involving noisy signal power and noise power of said audio signal.

39. A method, as claimed in claim 34, wherein said first parameter comprises pitch gain, wherein said plurality of parameters further comprise codebook gain, wherein said generating comprises performing said plurality of decoding steps by generating a codebook vector, scaling said codebook vector by said codebook gain to generate a scaled codebook vector, generating first samples based on pitch period in response to said scaled codebook vector, scaling said first samples by said pitch gain to generate first

scaled samples, and modifying said pitch gain to modify the contribution of said first scaled samples in order to manage said noise characteristic.

40. A method, as claimed in claim 34, wherein said first parameter comprises pitch gain, wherein said plurality of characteristics further comprises signal to noise ratio, wherein said generating comprises generating said adjusted first parameter in response to said pitch gain and said signal to noise ratio, and wherein said adjusted first parameter comprises an adjusted pitch gain.

41. A method, as claimed in claim 40, wherein said signal to noise ratio comprises a ratio involving noisy signal power and noise power of said audio signal.

42. A method, as claimed in claim 34, wherein said first parameter comprises pitch gain, wherein said plurality of parameters further comprise codebook gain, wherein said generating comprises performing said plurality of decoding steps to generate a codebook vector, scaling said codebook vector by said codebook gain to generate a scaled codebook vector, generating a power signal representing the power of said scaled codebook vector, and generating said adjusted first parameter in response to said pitch gain and said power signal, and wherein said adjusted first parameter comprises an adjusted pitch gain.

43. A method, as claimed in claim 34, wherein said first parameter comprises pitch gain, wherein said generating comprises generating at least first samples based on pitch period, scaling said first samples by said pitch gain to generate at least first scaled samples, generating at least a first power signal representing the power of said first scaled samples, and generating said adjusted first parameter in response to at least said pitch gain and said first power signal, and wherein said adjusted first parameter comprises an adjusted pitch gain.

44. A method, as claimed in claim 43, wherein said generating further comprises generating second samples based on pitch period responsive in part to said first power signal, scaling said second samples by said pitch gain to generate second scaled samples, generating a second power signal representing the power of said second scaled samples and generating said adjusted first parameter in response to said pitch gain, said first power signal and said second power signal.

45. A method, as claimed in claim 44, wherein said system comprises one or more long-term predictor buffers and wherein said generating said first and second samples comprises using said one or more buffers.

46. A method, as claimed in claim 34, wherein said first parameter comprises pitch gain, wherein said plurality of parameters further comprises a codebook gain, and wherein said generating comprises performing said plurality of decoding steps to generate a first vector, scaling said first vector by said codebook gain to generate a scaled codebook vector, filtering said scaled codebook vector by pitch synthesis filtering to generate a second vector, generating a power signal representing the power of said second vector, and generating said adjusted first parameter in response to said pitch gain and said power signal, and wherein said adjusted first parameter comprises an adjusted pitch gain.

47. A method, as claimed in claim 46, wherein said first vector comprises a codebook excitation vector and wherein said second vector comprises an LPC excitation vector.

48. A method, as claimed in claim 34, wherein said first parameter comprises a codebook vector comprising pulses using variable sets of amplitudes, wherein said generating comprises analyzing said sets to identify the powers of said noise

characteristic represented by said sets, identifying a first set representing a power less than the power represented by said sets other than said first set, and adjusting said pulses according to said first set to generate said adjusted parameter.

49. A method, as claimed in claim 34, wherein said plurality of decoding steps further comprises at least one decoding step that does not substantially affect the management of the noise characteristic and wherein said generating avoids performing said at least one decoding step.

50. A method, as claimed in claim 49, wherein said at least one decoding step comprises post-filtering.

51. A method, as claimed in claim 34, wherein said compression code comprises a linear predictive code.

52. A method, as claimed in -claim 34, wherein said compression code comprises regular pulse excitation – long term prediction code.

53. A method, as claimed in claim 34, wherein said compression code comprises code-excited linear prediction code.

54. A method, as claimed in claim 34, wherein said first parameter is a quantized first parameter and wherein said generating comprises generating said adjusted first parameter in part by quantizing said adjusted first parameter before replacing said first parameter with said adjusted first parameter.

55. A method, as claimed in claim 34, wherein said compression code is arranged in frames of said digital signals and wherein said frames comprise a plurality of subframes each comprising said first parameter, wherein said reading comprises reading at least said first parameter from each of said plurality of subframes in response to said

compression code, and wherein said replacing comprises replacing said first parameter with said adjusted first parameter in each of said plurality of subframes.

56. A method, as claimed in claim 55, wherein said replacing comprises replacing said first parameter with said adjusted first parameter for a first subframe before processing a subframe following the first subframe to achieve lower delay.

57. A method, as claimed in claim 34, wherein said compression code is arranged in frames of said digital signals and wherein said frames comprise a plurality of subframes each comprising said first parameter, wherein said generating comprises beginning to perform said decoding steps during a first of said subframes to generate a plurality of said decoded signals, wherein said reading comprises reading said first parameter from a second of said subframes occurring subsequent to said first subframe, wherein said generating further comprises generating said adjusted first parameter in response to said decoded signals and said first parameter, and wherein said replacing comprises replacing said first parameter of said second subframe with said adjusted first parameter.

58. A method, as claimed in claim 34, wherein said generating comprises performing at least one of a plurality of said decoding steps to generate decoded signals in response to said compression code and generating said adjusted first parameter in response to said decoded signals and said first parameter.

59. A method, as claimed in claim 34, wherein said first parameter is selected from the group consisting of codebook vector, codebook gain, pitch gain and LPC coefficients representations, including line spectral pairs and line spectral frequencies.

60. A method, as claimed in claim 34, wherein said audio signals have spectral regions affected by said noise characteristic, wherein said first parameter comprises a

representation of LPC coefficients, and wherein said generating comprises determining said spectral regions affected by noise in response to said compression code and said representation and generating said adjusted first parameter to manage said noise characteristic in said regions, and wherein said adjusted first parameter comprises an adjusted representation of LPC coefficients.

61. A method, as claimed in claim 60, wherein said representation of LPC coefficients is selected from the group consisting of line spectral frequencies and log area ratios.

62. In a communication system for transmitting digital signals comprising code samples, said code samples comprising first bits using a compression code and second bits using a linear code, said code samples representing an audio signal, said audio signal having a plurality of audio characteristics including a noise characteristic, a method of managing the noise characteristic without decoding said compression code comprising:

adjusting said first bits and said second bits in response to said second bits whereby the noise characteristic in the digital signals is controlled.

63. A method, as claimed in claim 62, wherein said linear code comprises pulse code modulation (PCM) code.

64. A method, as claimed in claim 62, wherein said code samples conform to the tandem-free operation of the global system for mobile communications standard.

65. A method, as claimed in claim 62, wherein said first bits comprise the two least significant bits of said samples and wherein said second bits comprise the 6 most significant bits of said samples.

66. A method, as claimed in claim 65, wherein said 6 most significant bits comprise PCM code.

Fig. 1

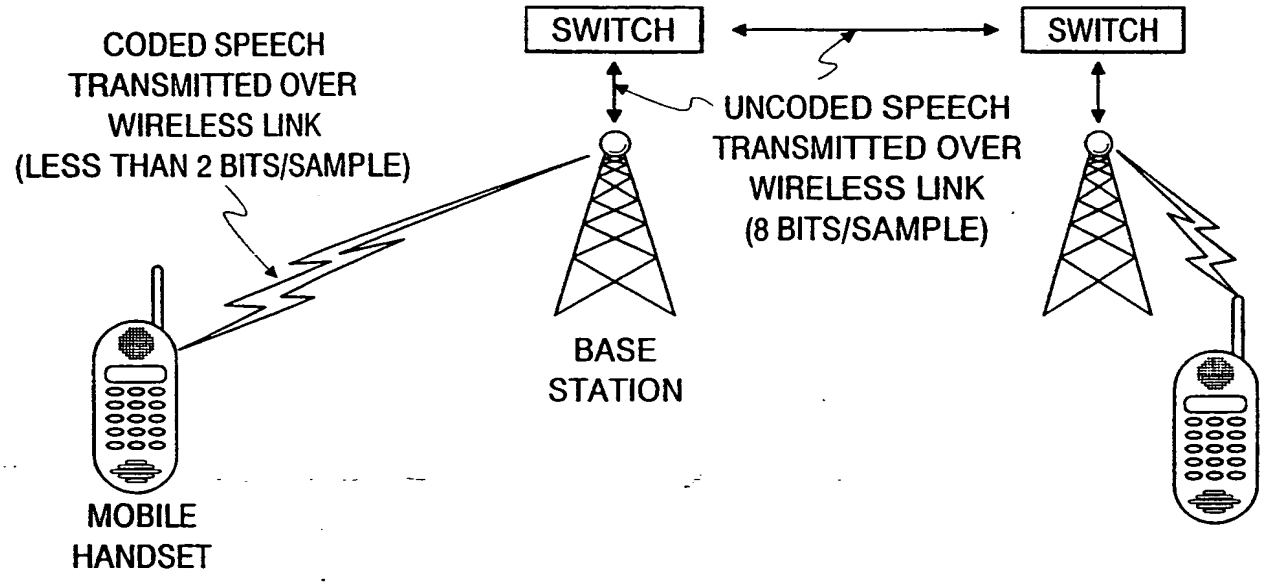
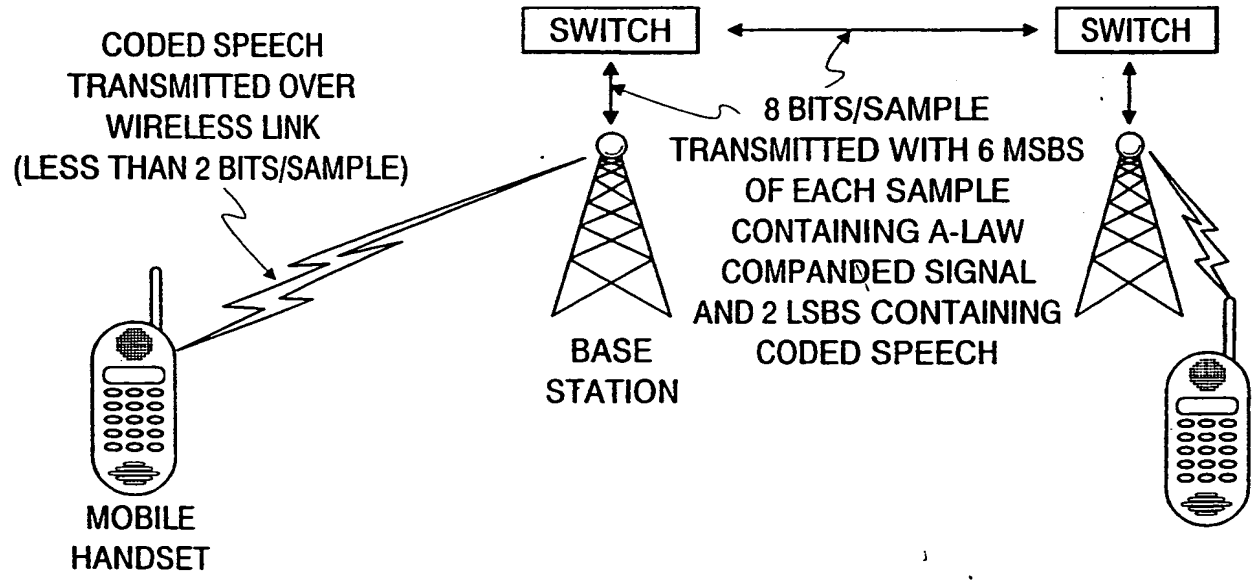


Fig. 2



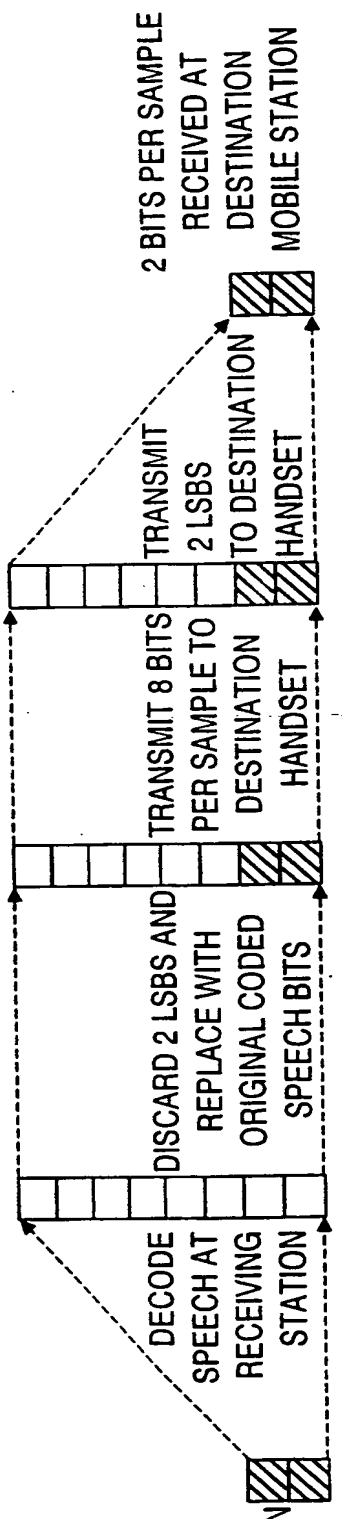


Fig. 3

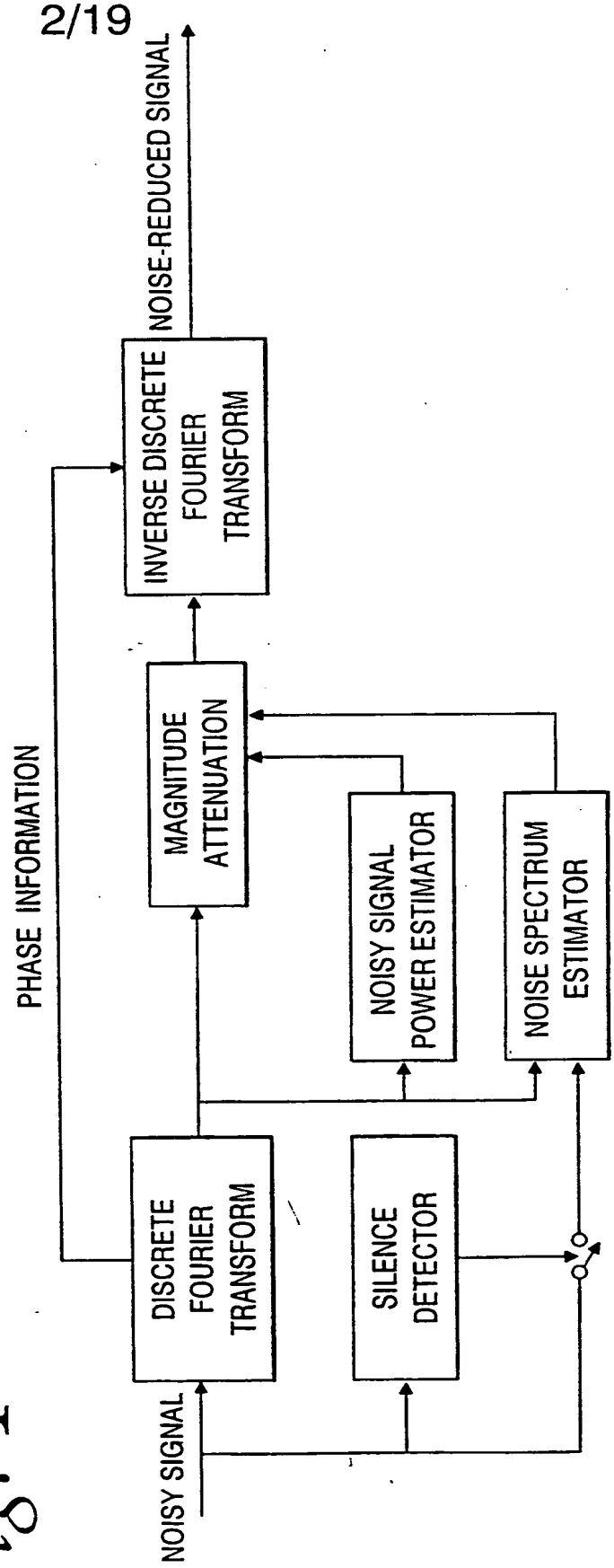


Fig. 4

Fig. 5

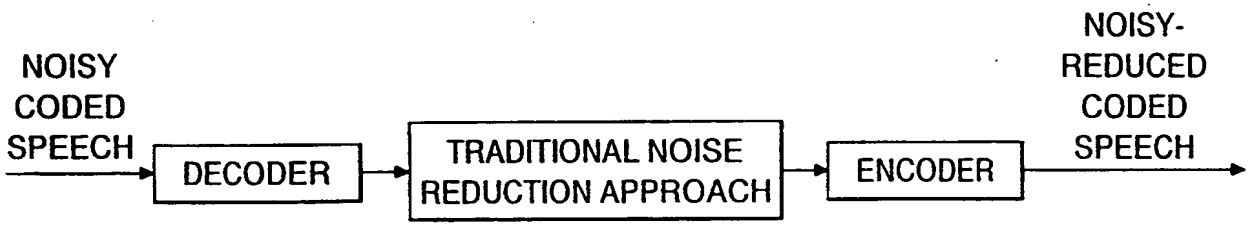


Fig. 6

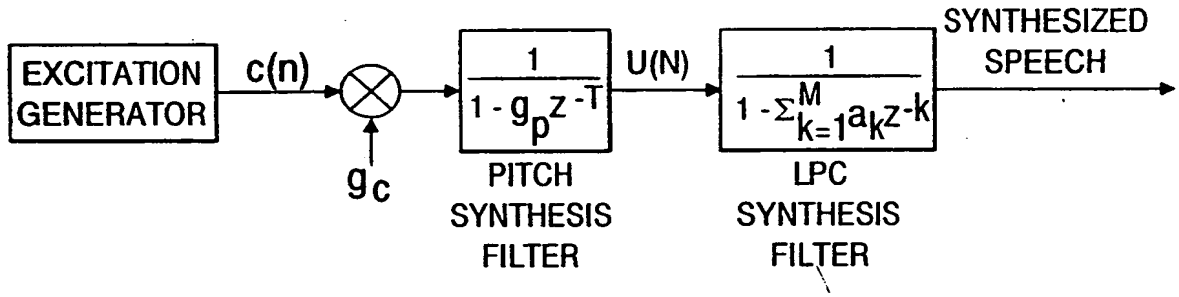


Fig. 7

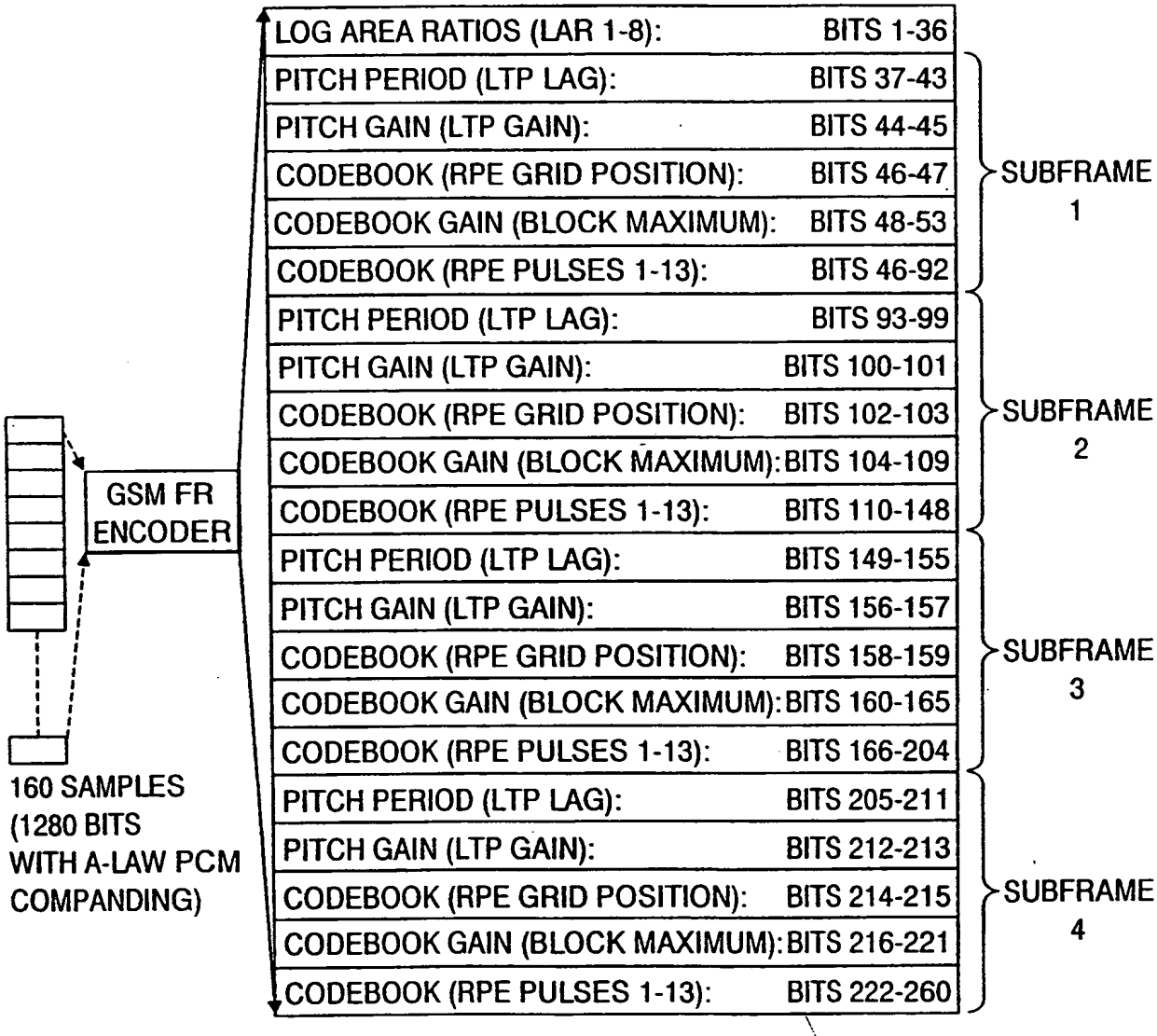


Fig. 8

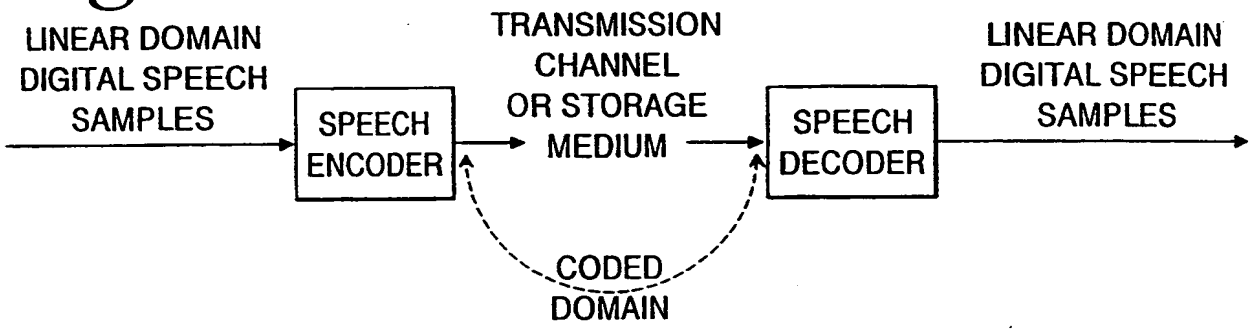


Fig. 9

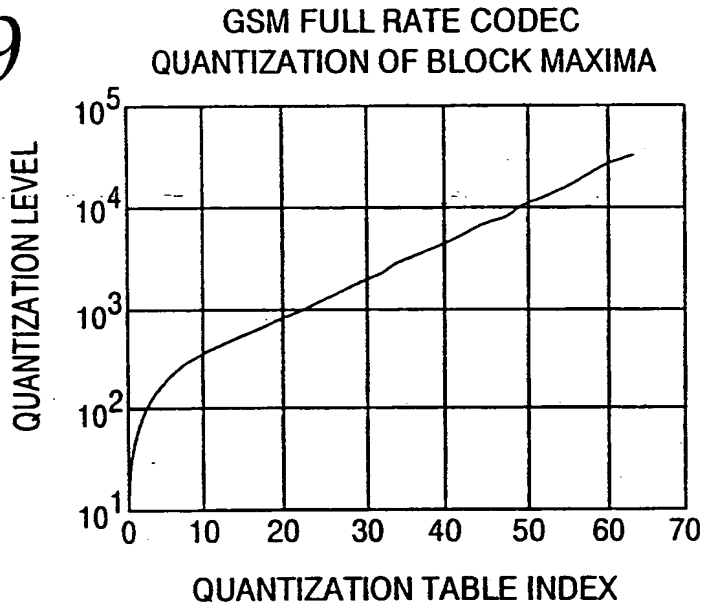


Fig. 10A

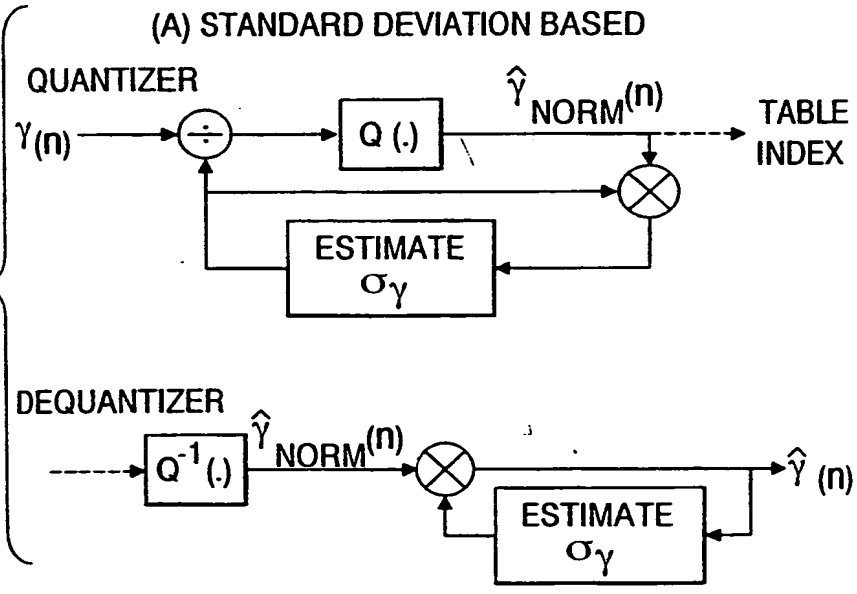


Fig. 10B

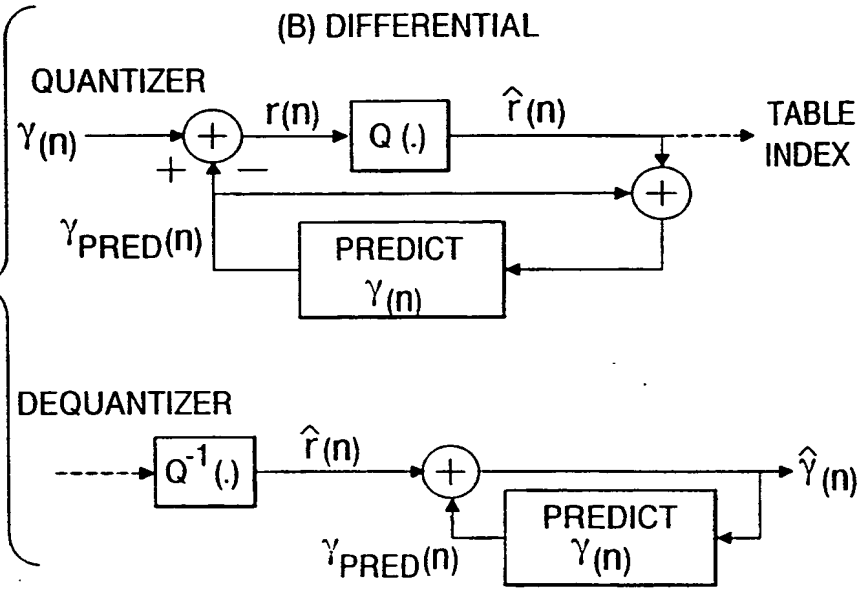


Fig. 11

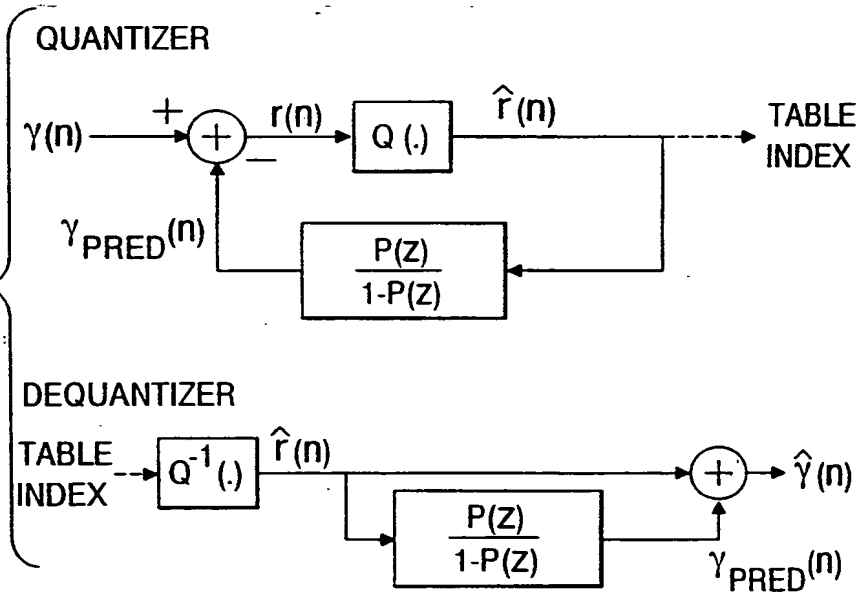


Fig. 12

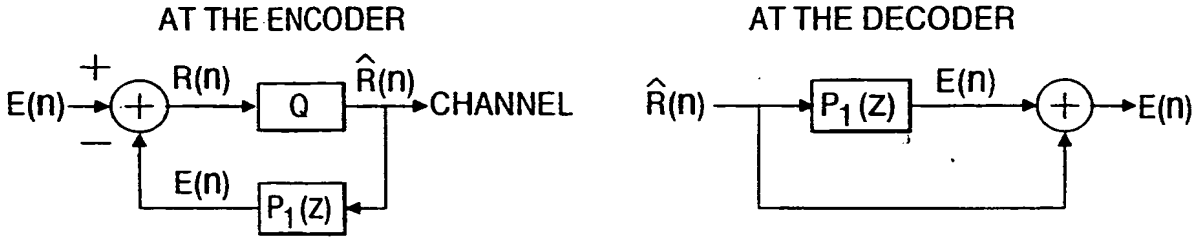


Fig. 13

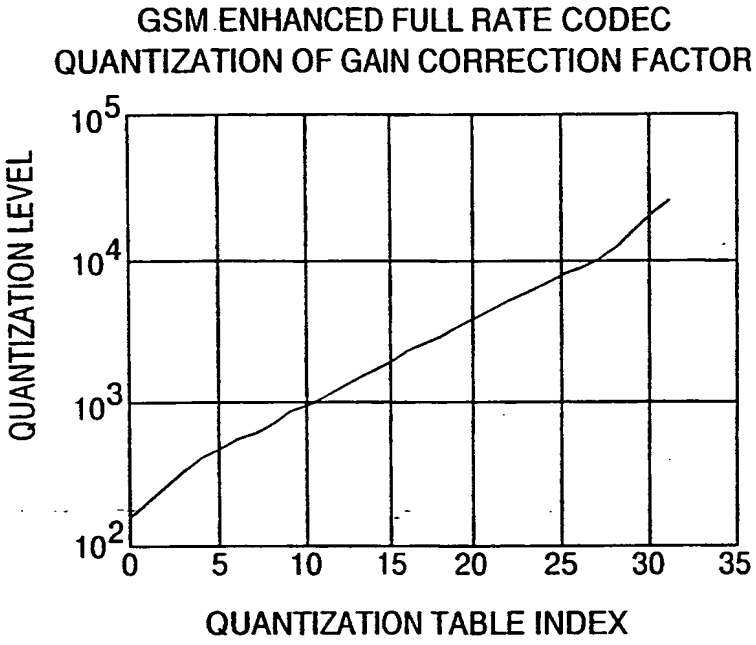


Fig. 14

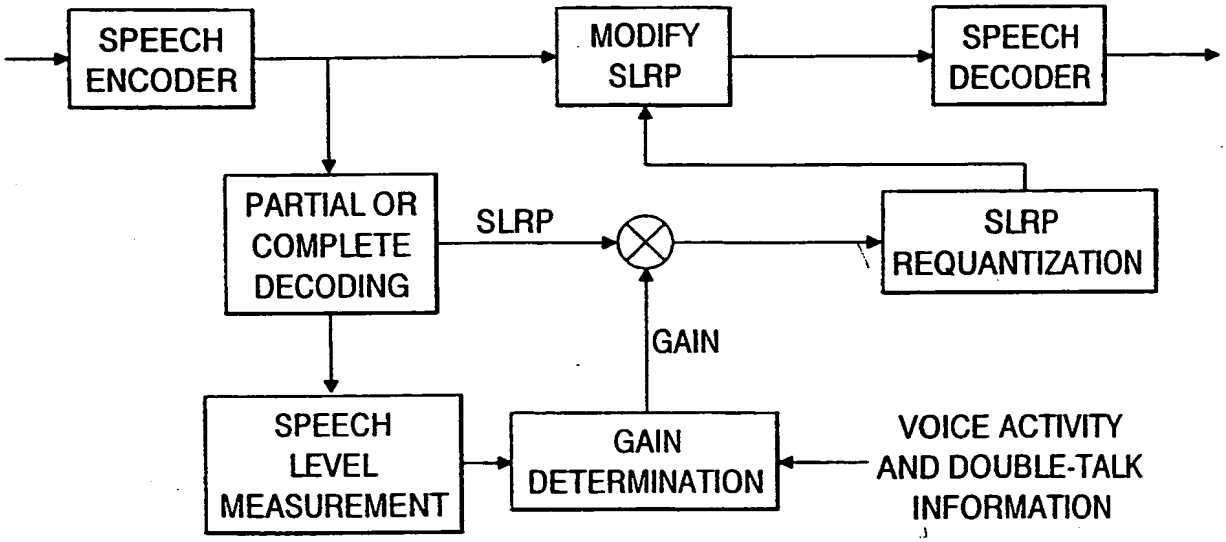


Fig. 15

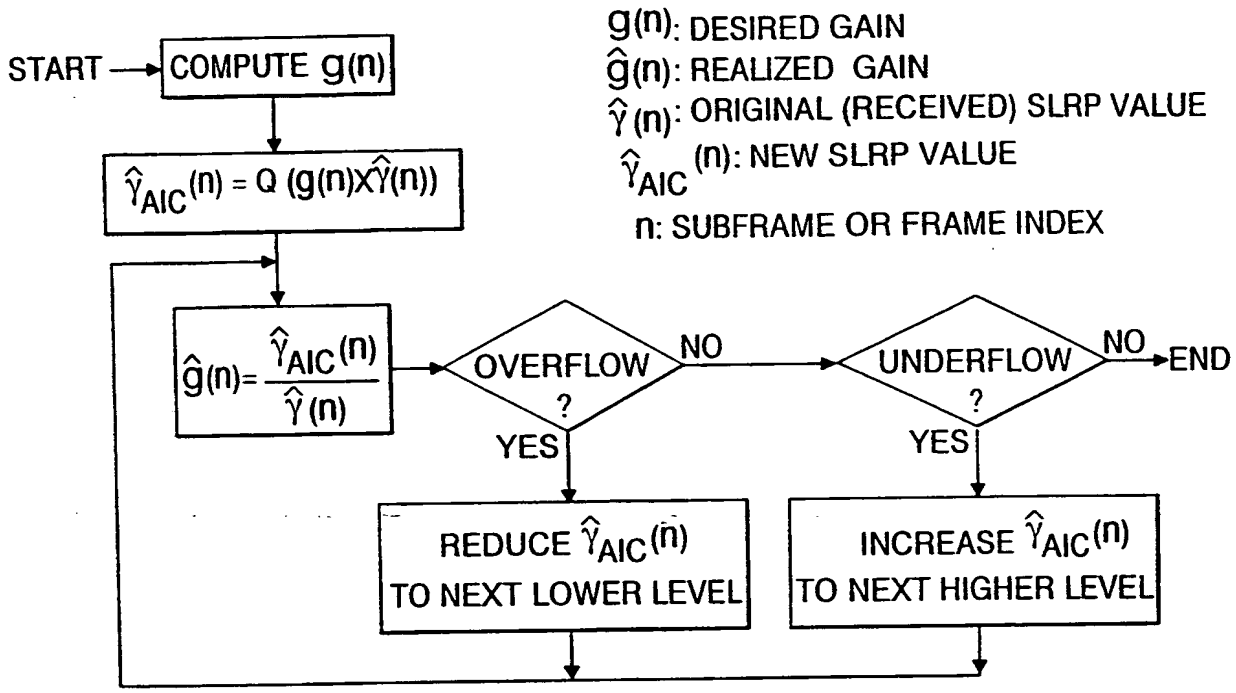


Fig. 16

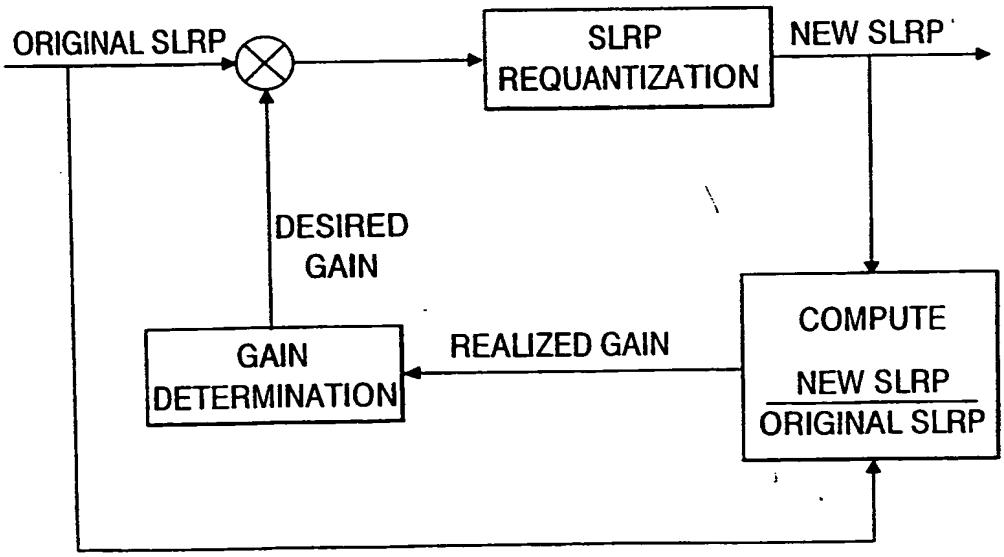
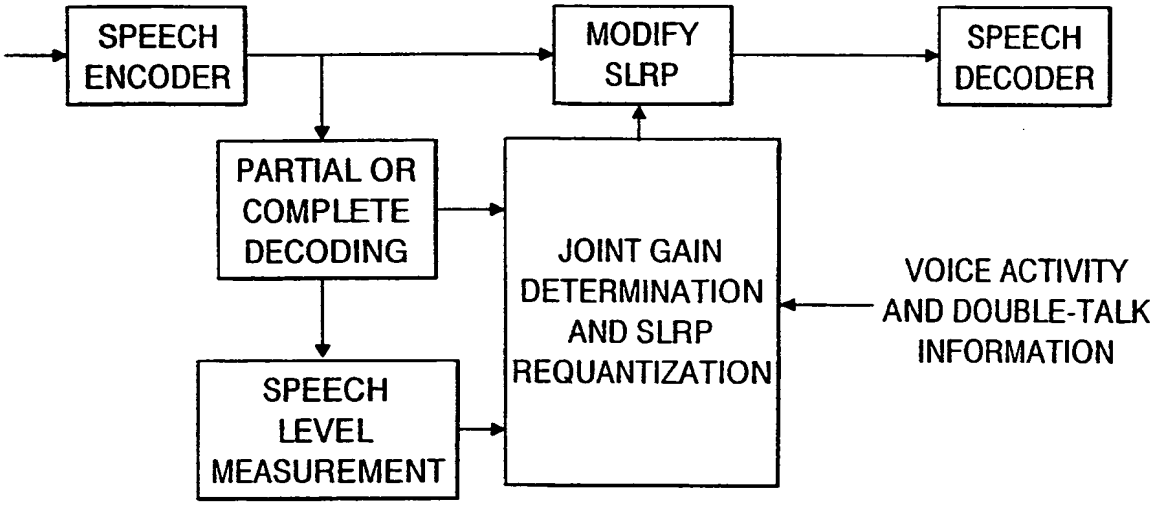


Fig. 17



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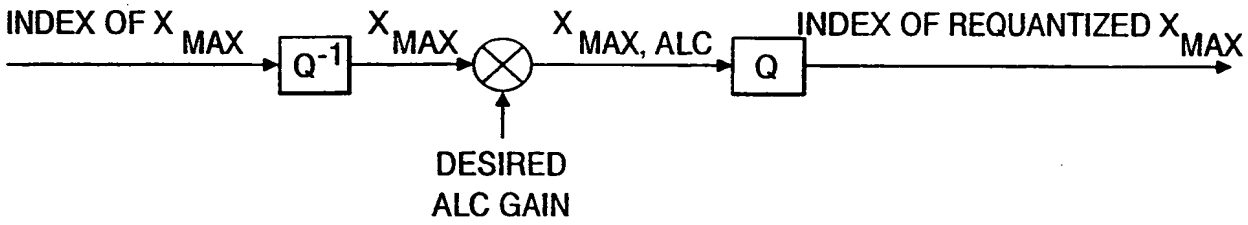


Fig. 19

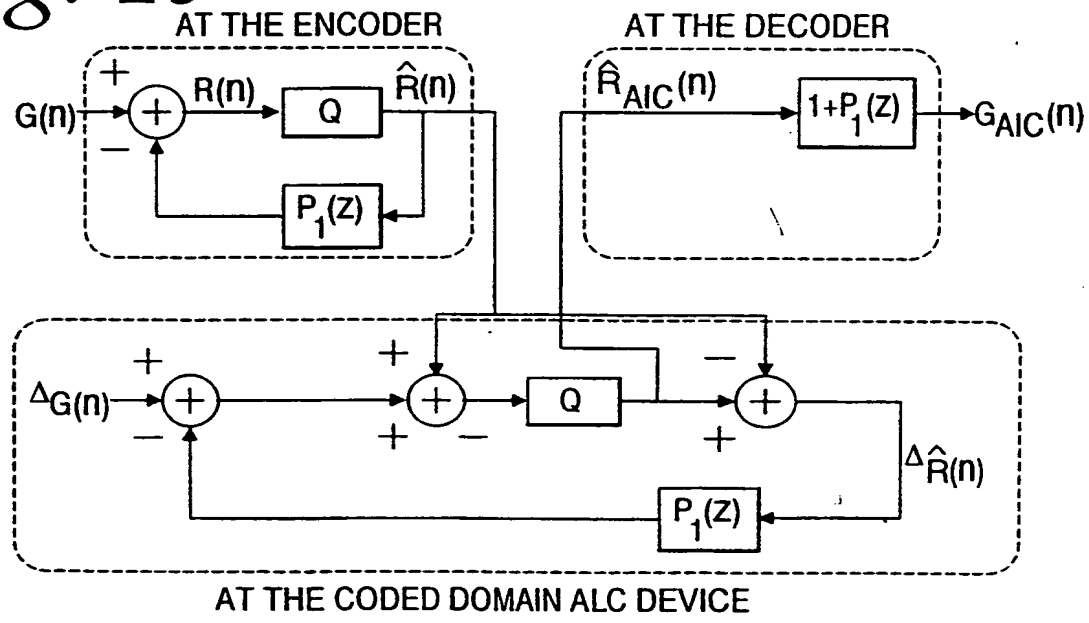


Fig. 20a

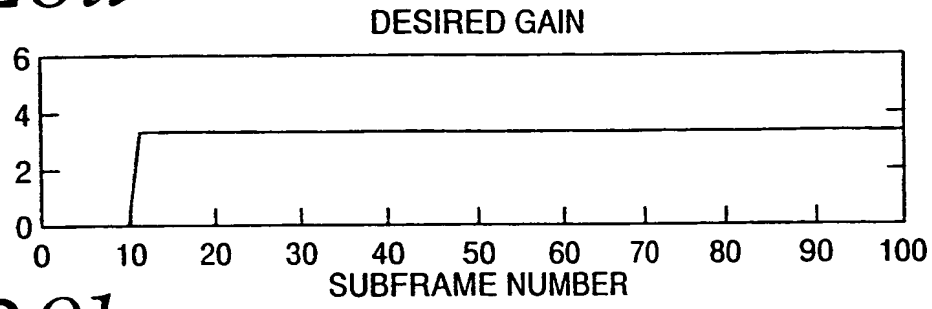


Fig. 20b

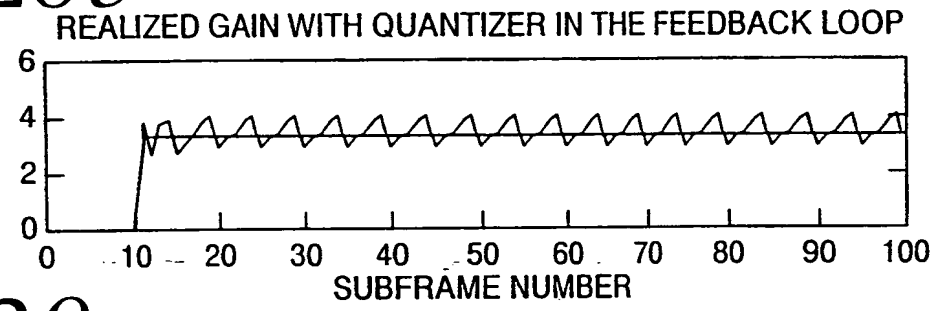


Fig. 20c

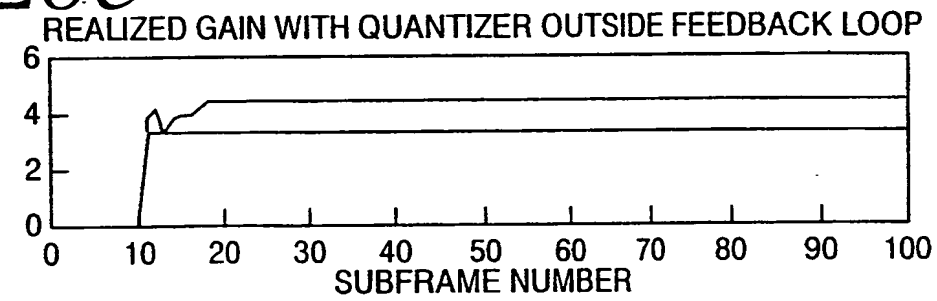


Fig. 21

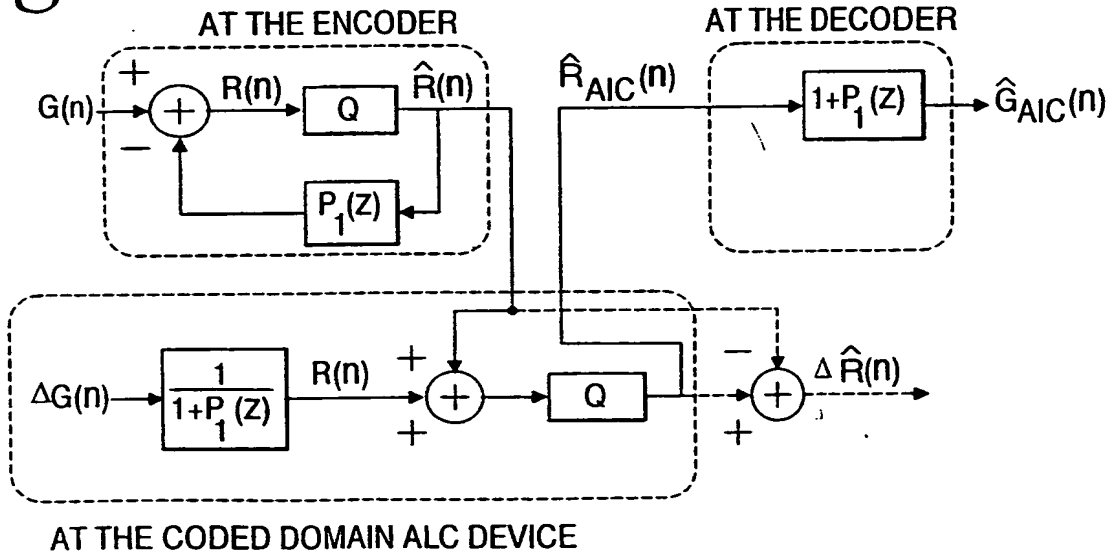


Fig. 22

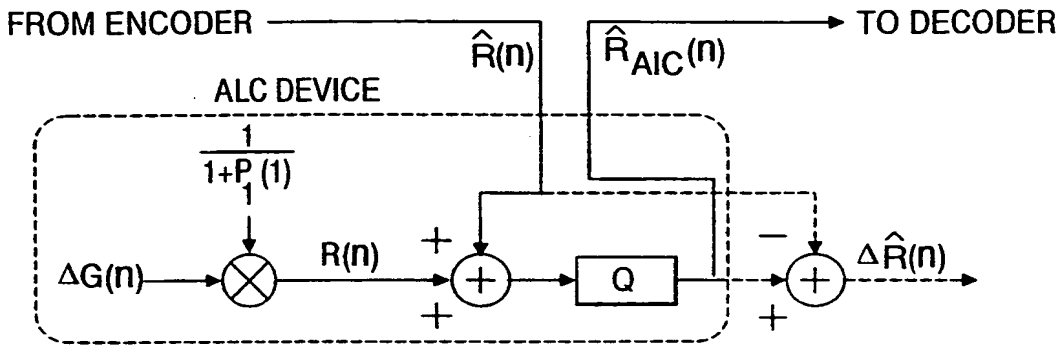


Fig. 23a

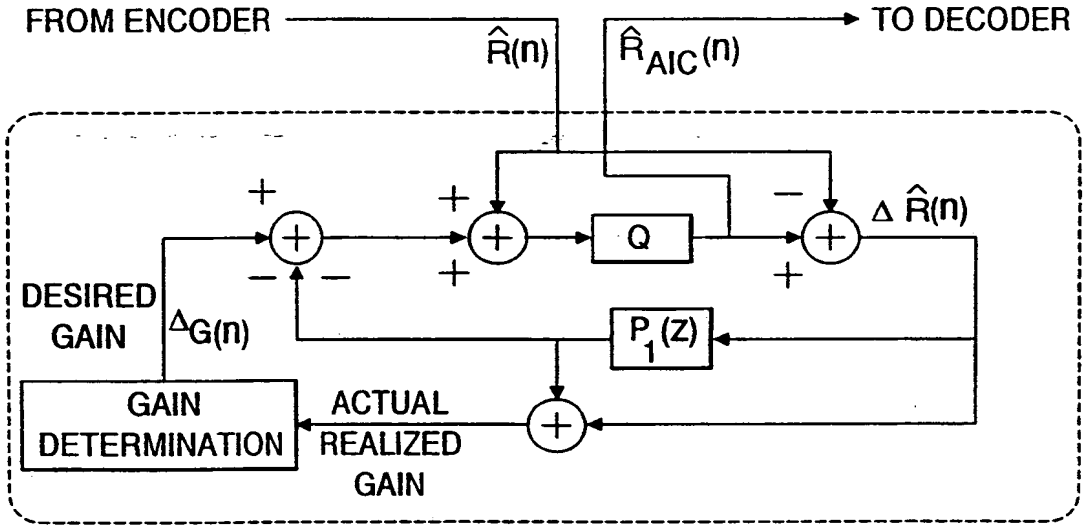


Fig. 23b

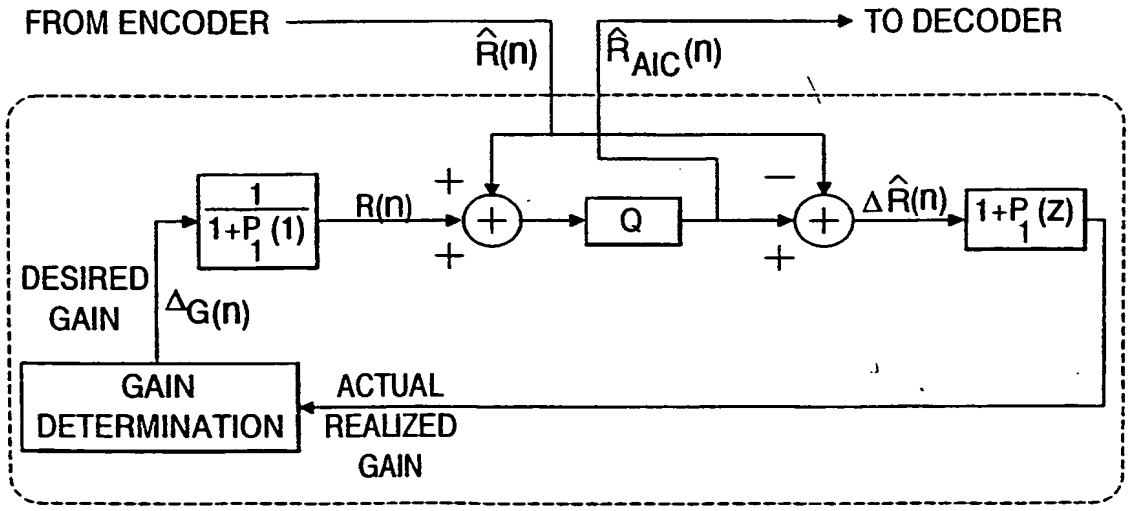


Fig. 24

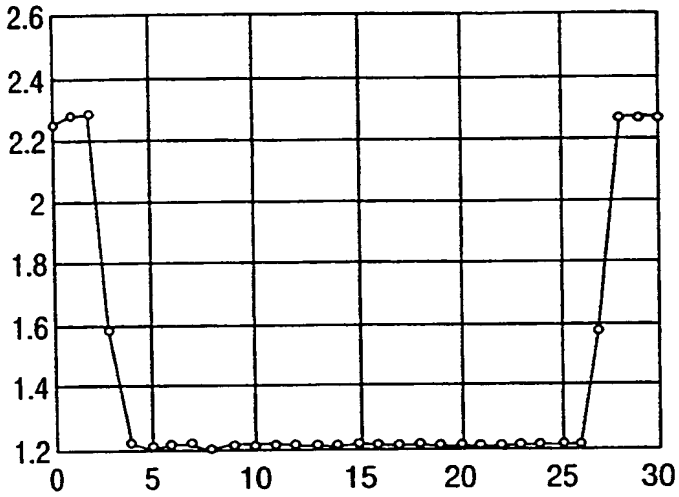


Fig. 25a

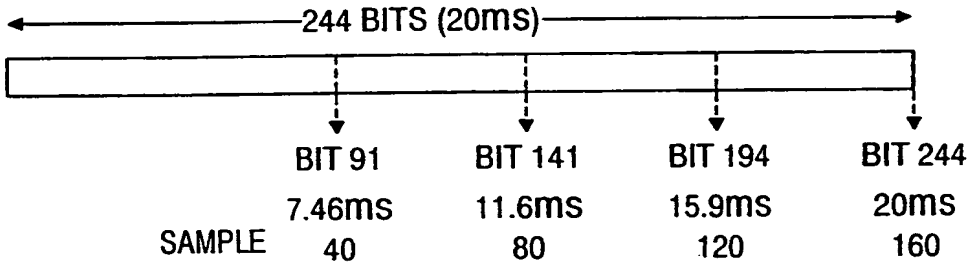
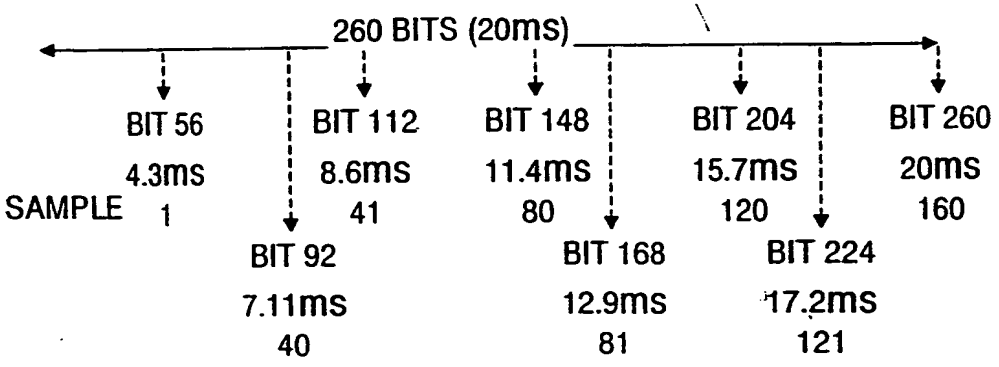


Fig. 25b



SINGLE-BAND LINEAR DOMAIN NOISE REDUCTION

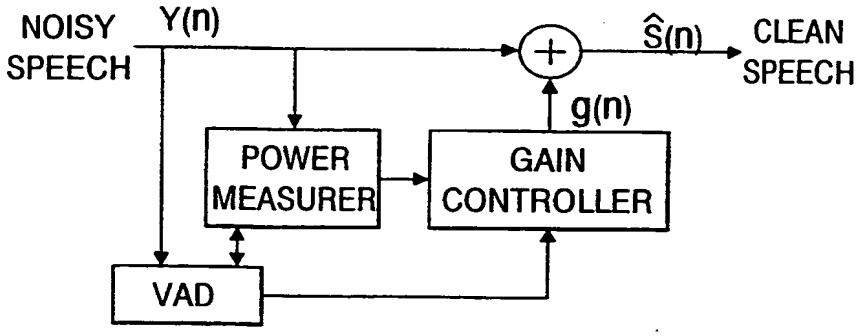


Fig. 26

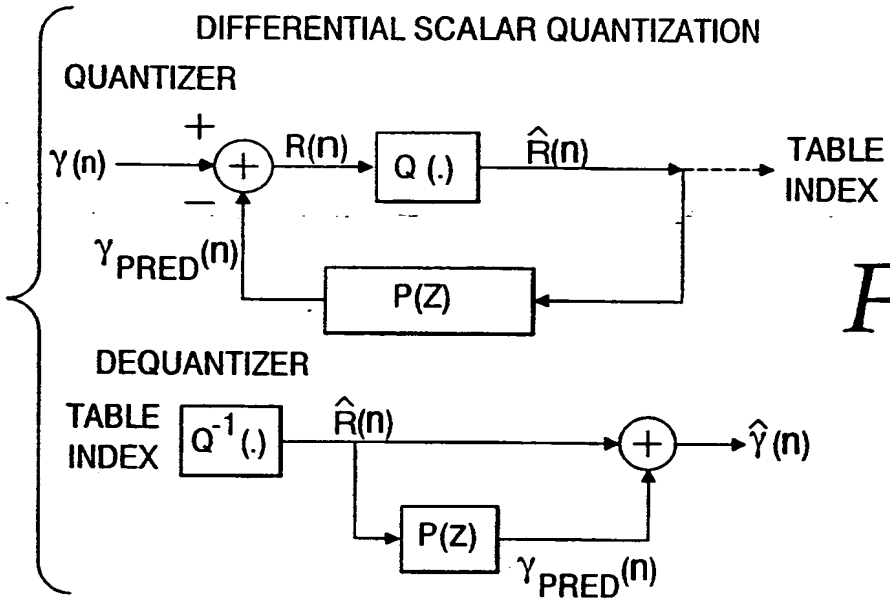
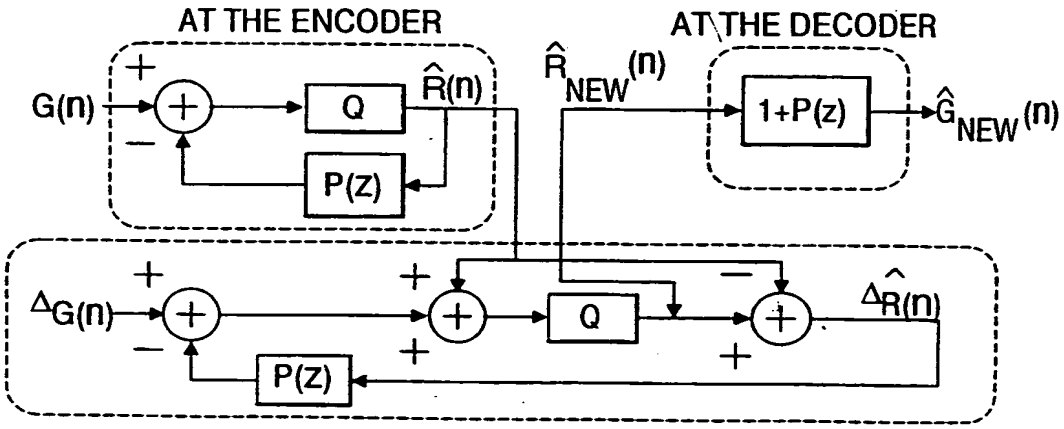


Fig. 27

Fig. 28

DIFFERENTIAL REQUANTIZATION OF A DIFFERENTIALLY QUANTIZED PARAMETER



AT THE NETWORK SPEECH ENHANCEMENT DEVICE

Fig. 29a

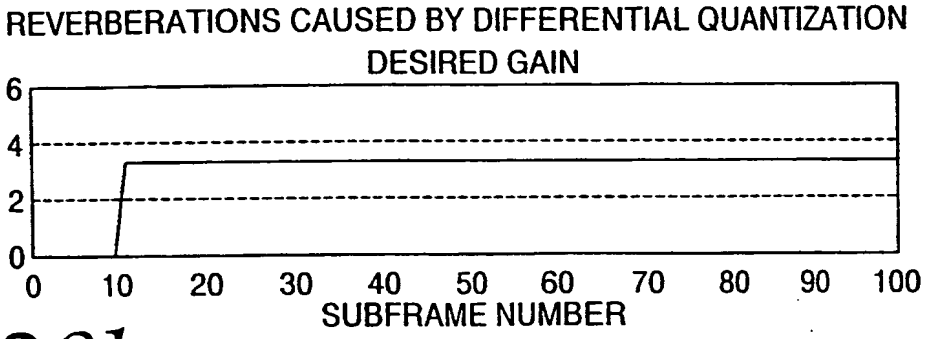


Fig. 29b

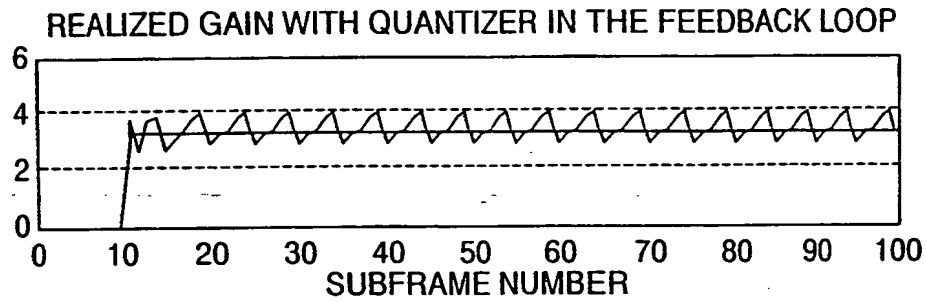


Fig. 29c

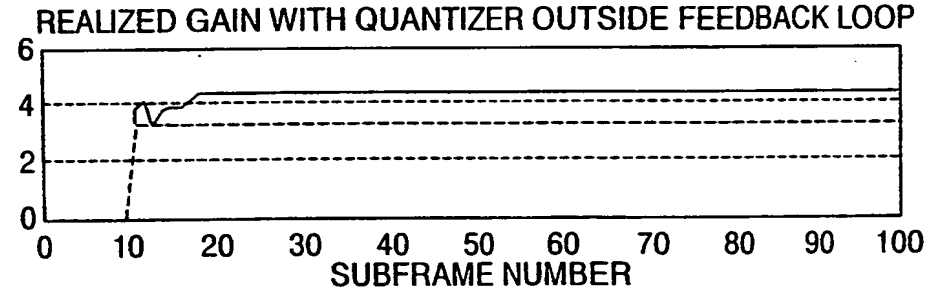


Fig. 30

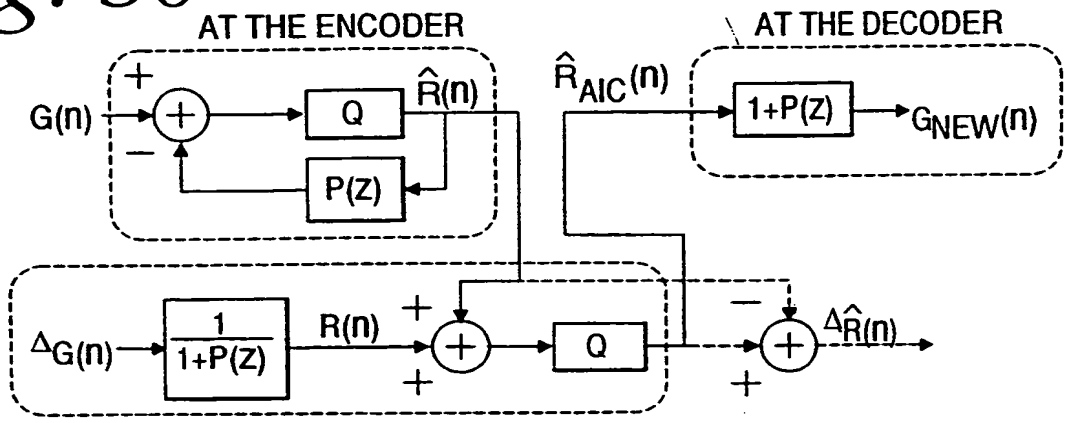


Fig. 31

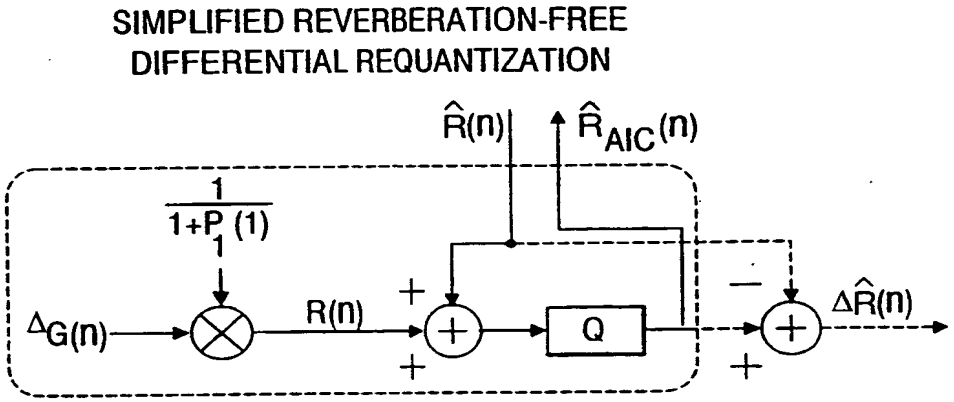


Fig. 32

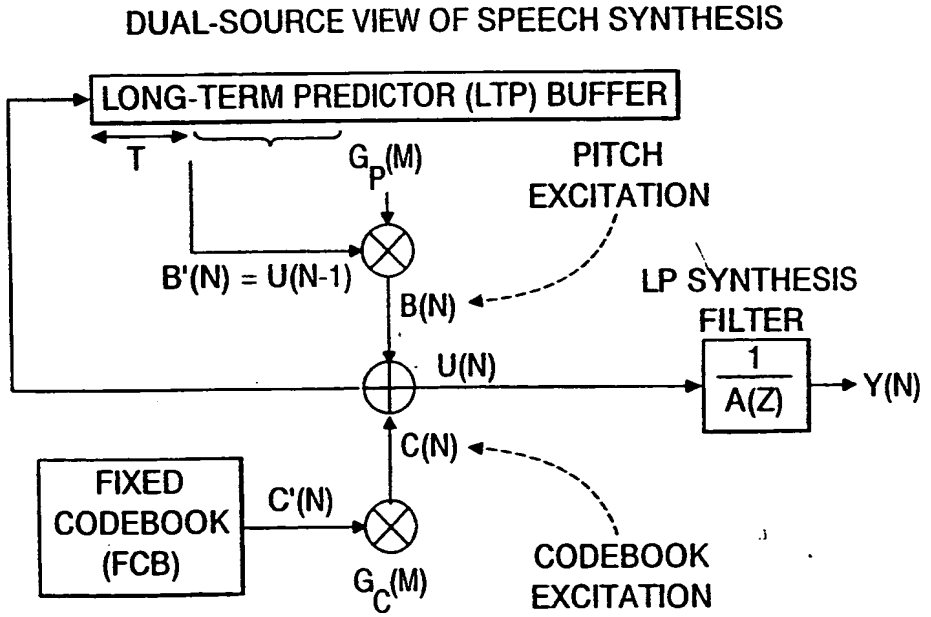


Fig. 33

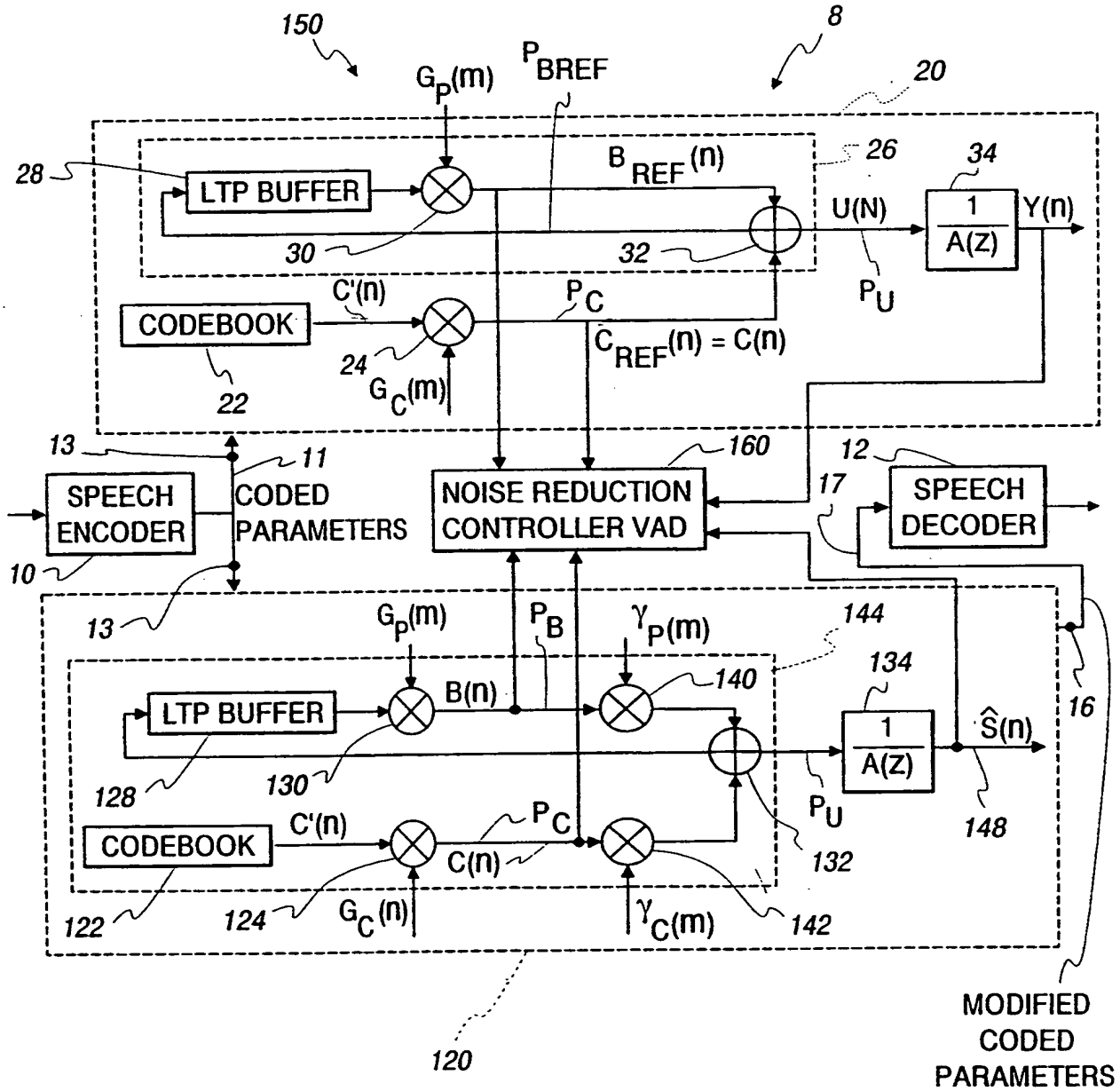


Fig. 34

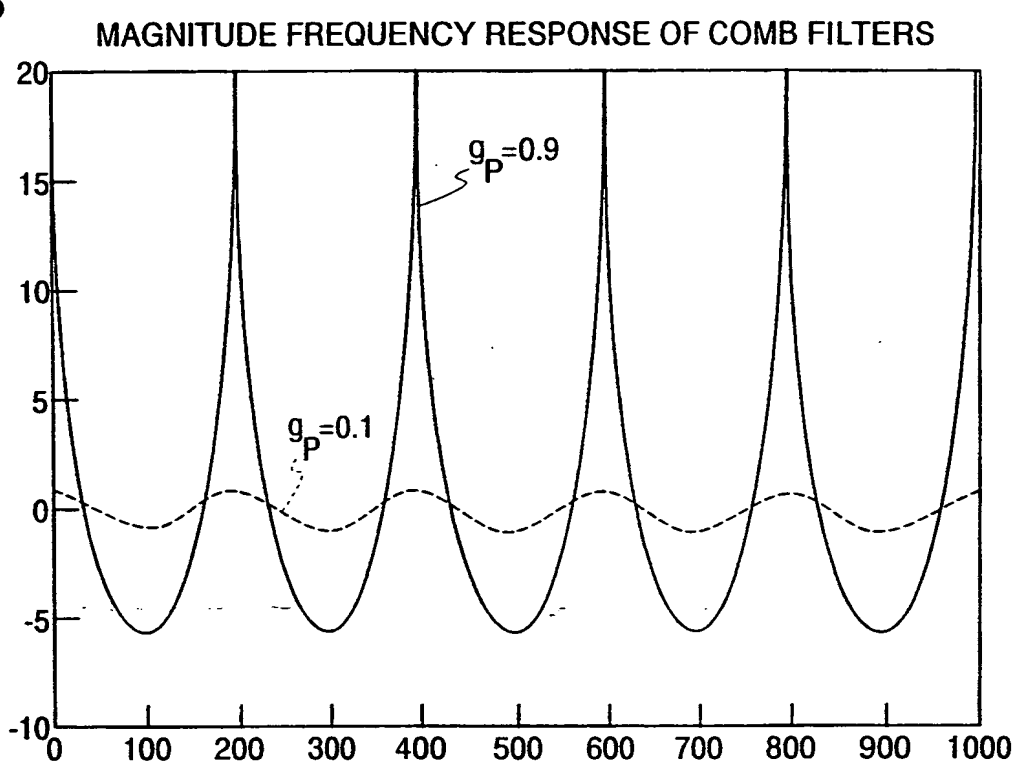


Fig. 35

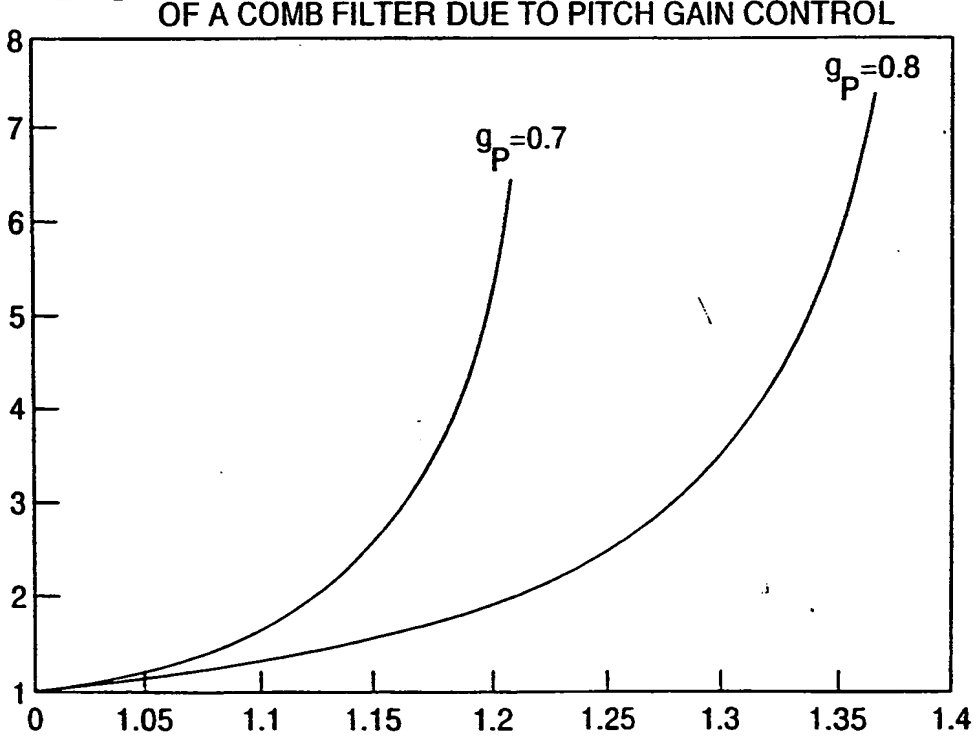


Fig. 36

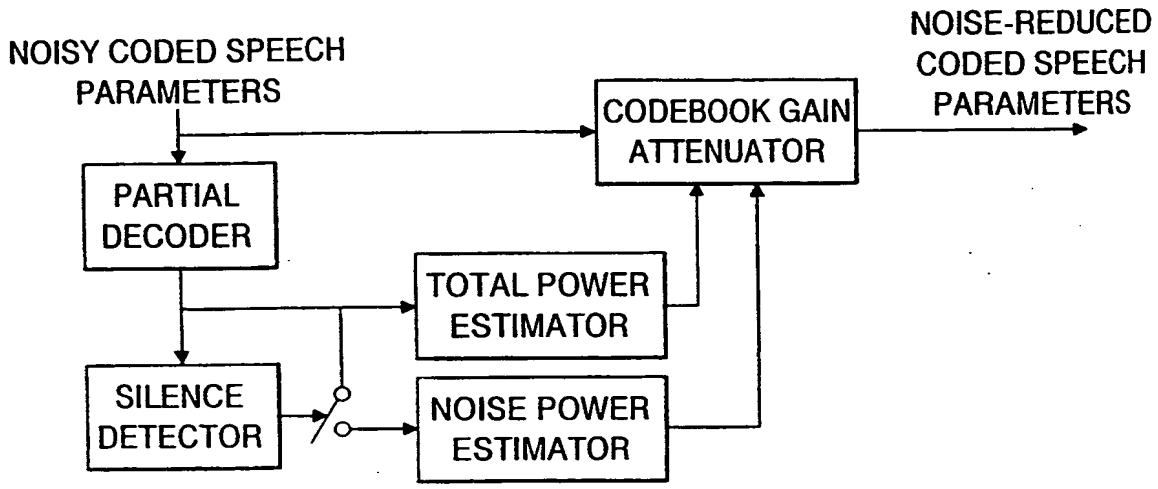


Fig. 37

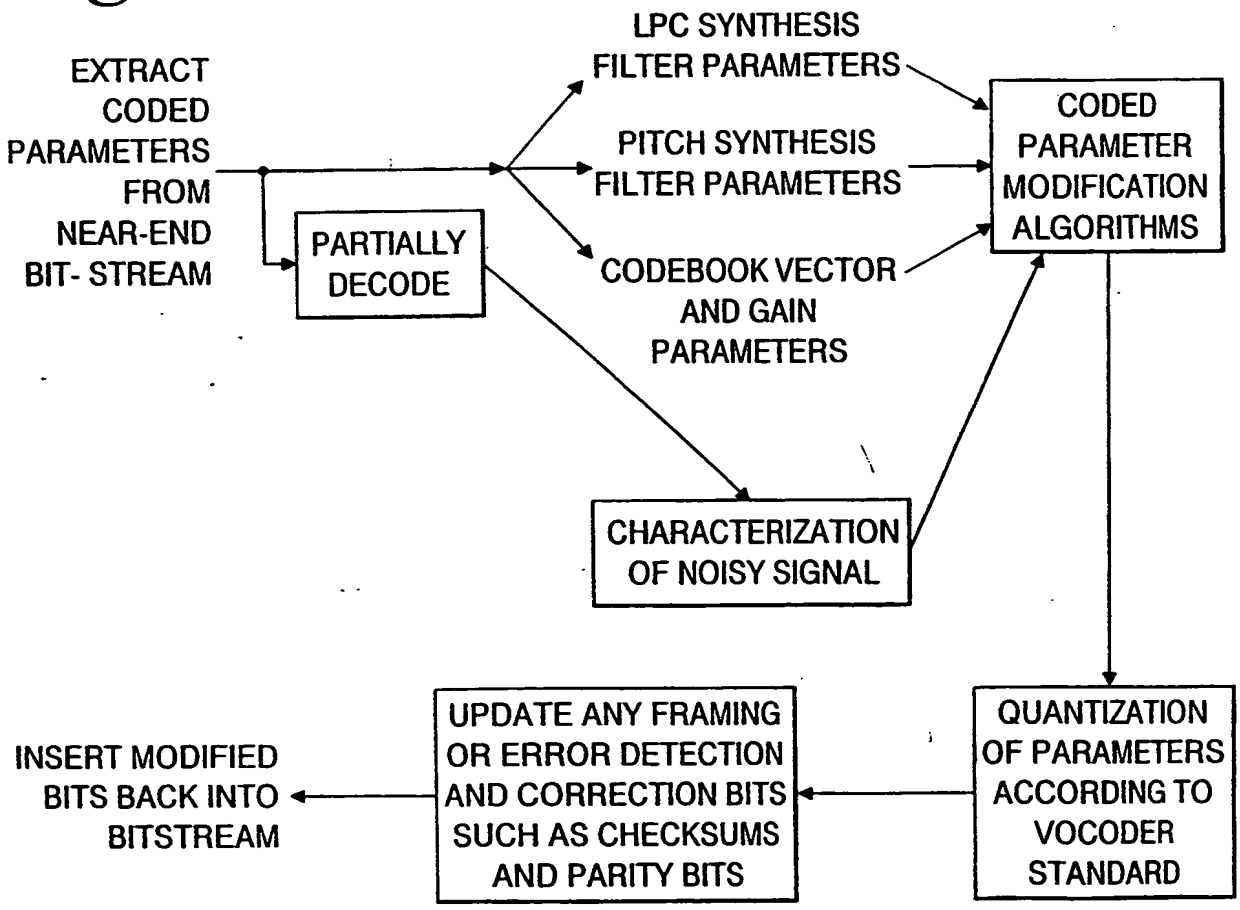


Fig. 38

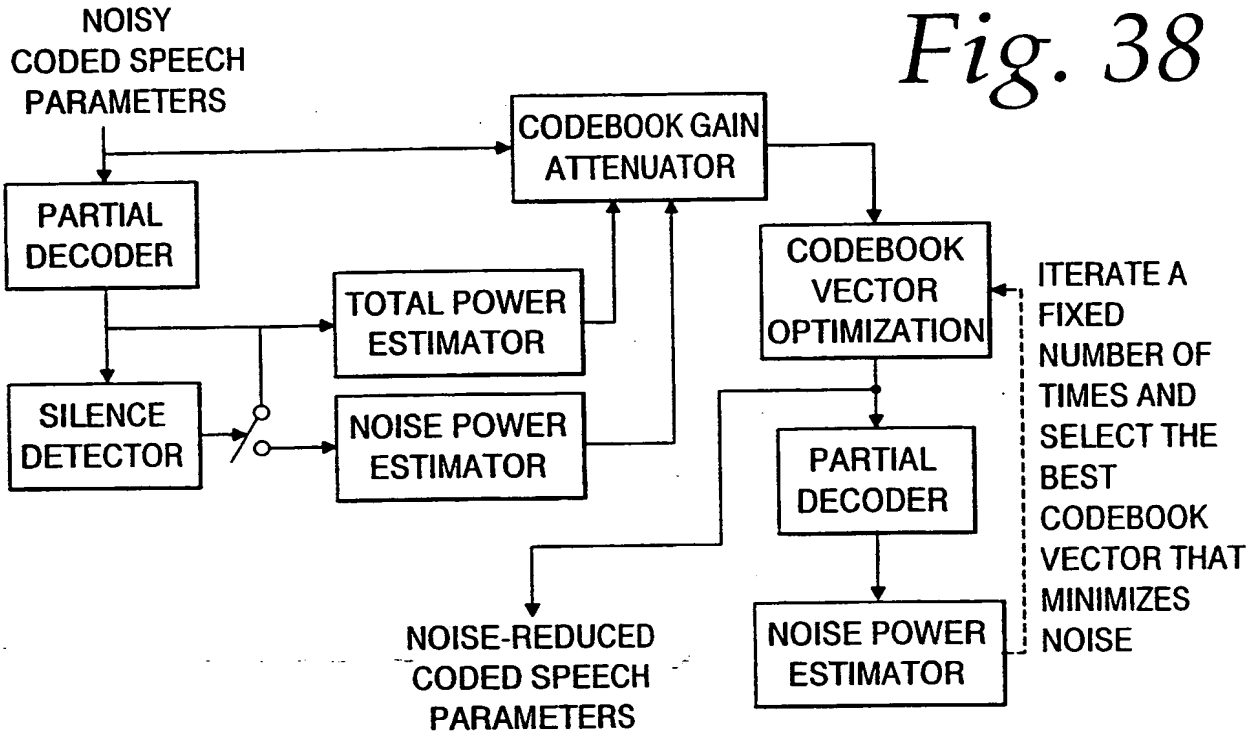
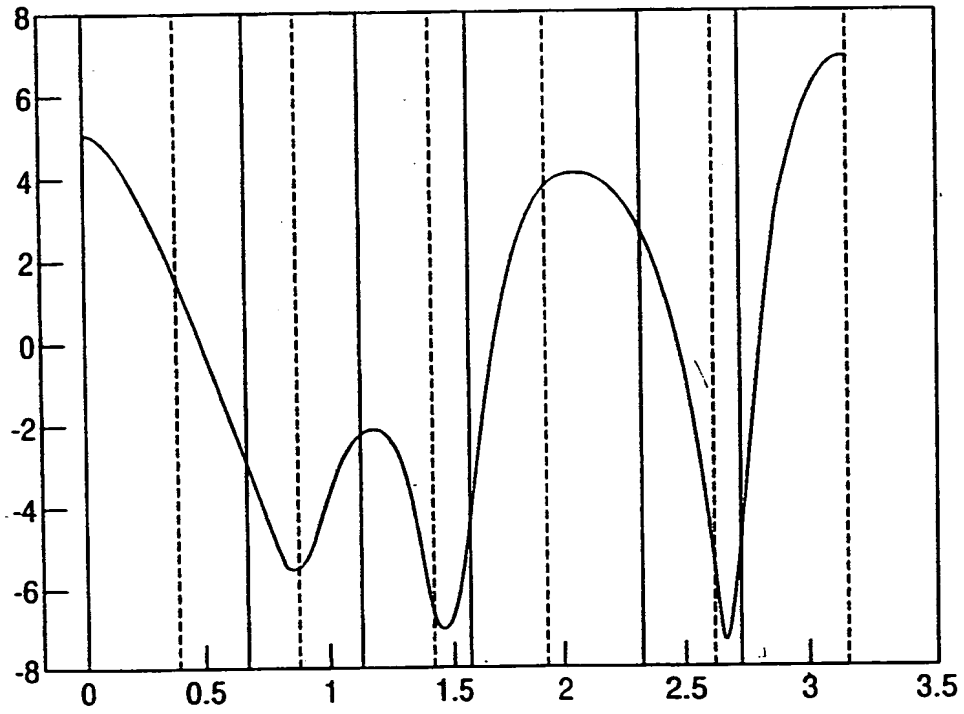


Fig. 39



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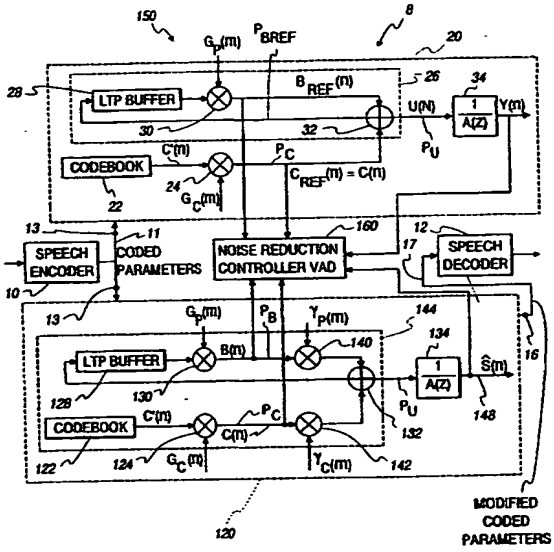
(54) Title: CODED DOMAIN NOISE CONTROL



(57) Abstract: A communications system (8) transmits digital signals using a compression code comprising a plurality of parameters including a first parameter. The parameters represent an audio signal comprising a plurality of audio characteristics, including a noise characteristic. The compression code is decodable by a plurality of decoding steps. A processor (150) is responsive to the compression code to read at least the first parameter. Based on such signals, the processor adjusts the first parameter and writes the adjusted first parameter into the compression code. As a result, the noise condition is effectively managed.

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US00/18165

A. CLASSIFICATION OF SUBJECT MATTER

IPC(7) :G10L 21/02
US CL :704/226, 230

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

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Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

IEEE DOCUMENTS

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

IEEE, WEST

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 5,097,507 A (ZINSER et al) 17 March 1992, Figs. 1,4;col. 2 line 44 - col. 3 line 5;col. 3 line 40 - col. 5 line 39; col. 5 line 55 - col. 7 line 34	1-66
X	US 4,969,192 A (CHEN et al) 06 November 1990, Fig. 3, abstract	1-4
X	US 5,680,508 A (LIU et al) 21 October 1997, Fig. 1	1-4

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

*	Special categories of cited documents:	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"A"	document defining the general state of the art which is not considered to be of particular relevance	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
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"O"	document referring to an oral disclosure, use, exhibition or other means		
"P"	document published prior to the international filing date but later than the priority date claimed		

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